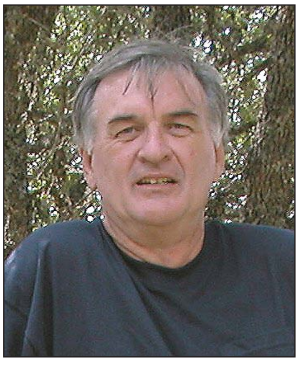
# CHAPTER 11

# Phased Arrays

Hardly any active ham needs an introduction to John Brosnahan, WØUN. Another American antenna guru, John retired as president of Alpha/Power, Inc, the Colorado-based manufacturer of top-notch power amplifiers. He is now relocating to the hill country of south-central Texas, moving all of his Colorado antennas and towers to his new 126-acre (50-hectare) site. Although the new site lacks a lot in ground conductivity, it should more than make up for that with its lack of man-made noise and no tower regulations! Although I had met John eye-to-eye on a number of occasions at Dayton Hamventions over the years it was during WRTC 1996 in San Francisco that we got to know each other better, and I have followed his moves in Amateur Radio ever since.

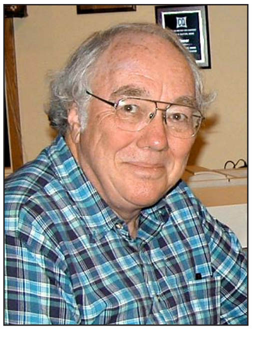


John is a research physicist by education who has spent his career on the electrical engineering side of remote-sensing instrumentation. From 1973 to 1978 he was the engineer for the University of Colorado's radio-astronomy observatory, designing receivers and antenna arrays for HF and VHF radio astronomy. Since then he has been founder and president of two companies that design and build HF and VHF radar systems for remote sensing of the atmosphere and ionosphere. He has designed and built arrays all the way from 80 dipoles at 2.66 MHz, which covered 40 acres (16 hectares), to a 12,288-dipole array at 49 MHz. He has also built numerous Yagi arrays, including a 768-element array at 52 MHz to a 500-element array at 404 MHz.

When I asked him, John immediately volunteered to review the chapters on arrays for the low bands, as he did in the previous edition. Thank you, John, for your help, your input and also your friendship!

When John had finished his proofreading he wrote to me: "*I loved this chapter. Excellent balance of the whole range of typical ham phased arrays, with a lot of very solid practical information and enough new stuff to make it worth the money to buy the new edition.*"

Corresponding with Robye Lahlum on the issue of L-networks for feeding arrays was more than interesting. Robye developed the mathematics for feeding the arrays in a Lewallen fashion, but with no limitation of phase angle. You can design a network for any array in a minute with the mathematics Robye provided for this book. Robye also developed a novel test setup that allows us to adjust the components of the L-network until we are right on the nose.



In addition, Robye has proven to be a very meticulous and thorough proofreader! I am proud having your contributions in my new book, *Robye.*

John Battin , K9DX, was the first to dare building a 9-Circle array, which I described in the Third Edition of this book. It's been a very enlightening experience for me discussing various issues of array feeding with John, and I am extremely grateful to him for letting me share his experience with the readers of this book. Thank you John.

If you want gain and directivity on one of the low bands and if you live in an area with good or excellent ground, an array made of vertical elements may be the answer, provided you have room for it. Arrays made with vertical elements have the same requirements as single vertical antennas so far as ground quality is concerned. Before you decide to put one up, take the time to understand the mechanism of an array with all-fed elements.

In this chapter I cover the subject of arrays made of elements that, by themselves, have an omnidirectional horizontal radiation pattern; that is, vertical antennas.

## 1. RADIATION PATTERNS

### 1.1. How the Pattern is Formed

In Chapter 7 we explored in great detail how the radiation pattern of an array was formed.

### 1.2. Directivity Over Perfect Ground

**Fig 11-2** shows a range of radiation patterns obtained by different combinations of two monopoles over perfect ground and at a 0° elevation wave angle. These directivity patterns are classics in every good antenna handbook.

### 1.3. Directivity Over Real Ground

Over real ground there is no radiation at a 0° elevation angle. All the effects of real ground, which were described in detail in Chapter 9 on verticals, apply to arrays of verticals.

### 1.4. Direction of Firing

The rule is simple: An array always fires in the direction of the element with the lagging feed current.

### 1.5. Phase Angle Sign

Phase angles are a relative thing, which means you can put your *reference* phase angle of 0° anywhere in the array. We will stick to our own convention of assigning the 0° phase angle to the back element of an array. This means that the feed currents in all other elements will carry a negative sign.

## 2. ARRAY ELEMENTS

In principle, you can use verticals of length longer than $\lambda/4$ (electrically) for building arrays, but in that case the various feed systems described in this chapter do not apply. However, the whole range of verticals described in Chapter 9 can be used as elements for these vertical arrays, provided they are base-fed and are not longer than $\lambda/4$ electrically.

Quarter-wave elements have gained a reputation for giving a reasonable match to a 50-Ω line, which is certainly true for single vertical antennas. In this chapter we will learn the reason why quarter-wave resonant verticals do not have a resistive 36-Ω feed-point impedance when operated in arrays (even assuming a perfect ground). Quarter-wave elements still remain a good choice, since they have a reasonably high radiation resistance. This ensures good overall efficiency. On 160 meters, the elements could be top-loaded verticals, as described in Chapter 9.

The design methodology for arrays given in Section 3, as well as all the designs described in Section 4, assume that all the array elements are physically identical, with a current distribution that is the same on each element. In practice this means that only elements with a length of up to $\lambda/4$ should be used. Remember, the patterns given in Section 4 do not apply if you use elements much longer than $\lambda/4$. They certainly do not apply for elements that are $\lambda/2$ or $5\lambda/8$ long. If you want to use long elements, you will have to model the design using the particular element lengths (Ref 959). This may be a problem if you want to use shunt-fed towers carrying HF beams as elements for an array. With their top loads, these towers are electrically often much longer than $\lambda/4$.

## 3. DESIGNING AN ARRAY

The radiation patterns shown in Fig 11-2 give a good idea what can be obtained with different spacings and different phase delays for a 2-element array. For arrays with more elements there are a number of popular classic designs. Many of those are covered in detail in this chapter. A good array should meet the following specifications:

- High gain (you want to be loud)
- Good directivity (F/B, forward beamwidth, especially if you will not be using separate receiving antennas)
- Ease of feeding
- Ease of direction switching

### 3.1. Modeling Arrays

If you feed the elements of an array at a current maximum (which is usually the case with base-fed elements not longer than $\lambda/4$ long) the RF drive has to be specified as a current. If you feed at a voltage maximum (which we hardly ever do in vertical arrays), the RF drive will need to be specified as a voltage. We normally define currents for the RF sources since it is the current (magnitude and phase) in each element that determines the radiation pattern in such an array. Therefore currents, rather than voltages, should be specified. All modern modeling programs allow you to define "sources" as current sources or as voltage sources.

You can do initial pattern assessments using a *MININEC*-based program, although the latest versions of some programs no longer include such a computing engine (eg, *EZNEC 4.0*) because modern computers allow fast analysis with *NEC*-engines. Also, if you want to do some modeling that includes the influence of a radial system plus the influence of a poor ground, the full-blown *NEC* program is required. Studies on elevated radial systems and on buried radials require *NEC* (*NEC-3* or better yet, *NEC-4* if buried radials are involved).

In this chapter, the influence of the loss introduced by an imperfect radial system has been included in the form of an equivalent loss resistance in series with each element feed point (most model used 2 Ω).

### 3.2 About Polar and Rectangular Coordinates

We will be going into detail on various issues and aspects of arrays and will be talking impedances all the time. It's a good idea to review a few basics.

- **Complex impedance**: A complex impedance is an impedance consisting of a real part (resistive part) and an imaginary part (reactive part).
- **Complex number**: A complex impedance is represented by a complex number.
- **Complex number representation**: While a real number can be represented as a point on a line, a complex number

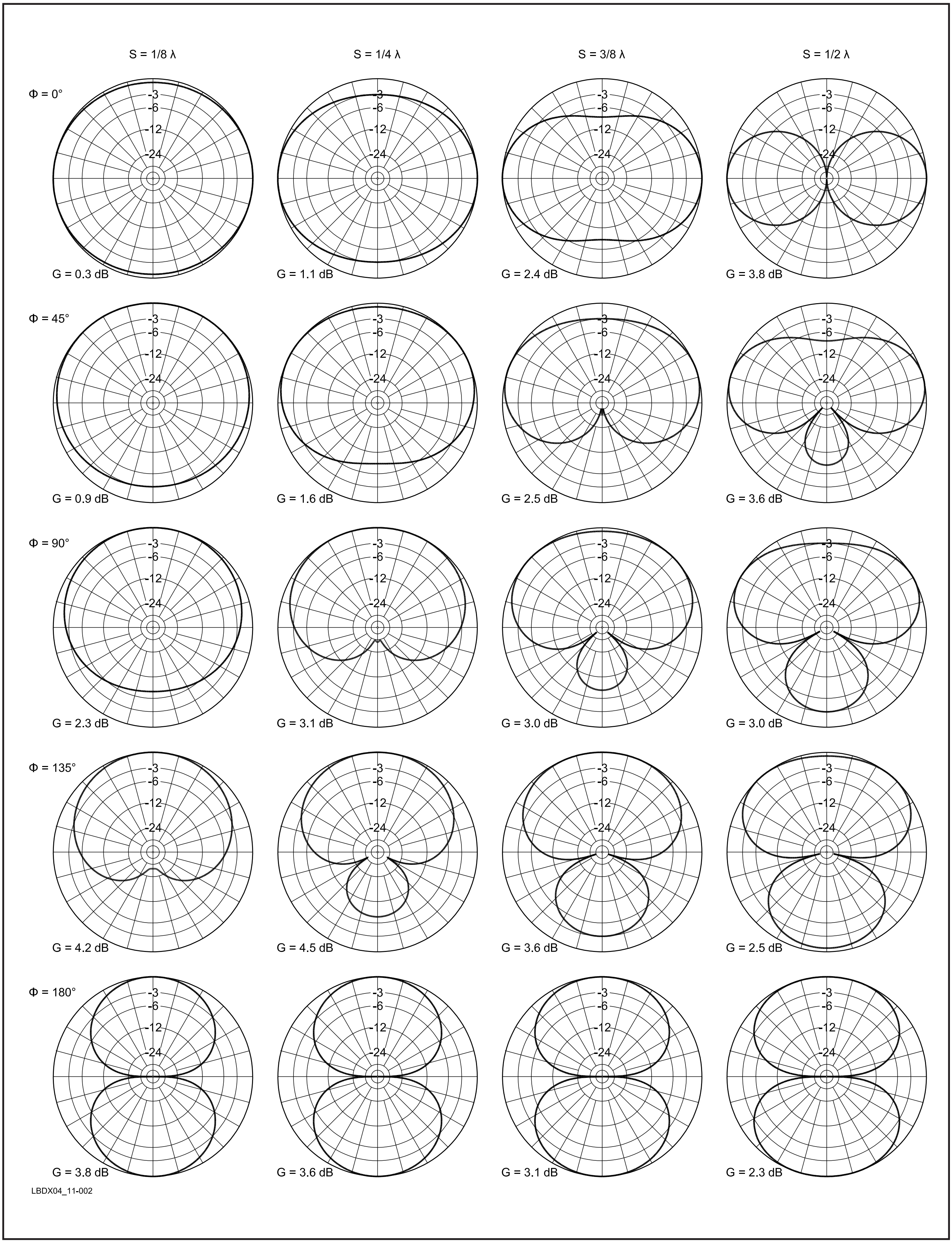



**Fig 11-2—Horizontal radiation patterns for 2-element vertical arrays (both elements fed with the same current magnitude). The elements are in the vertical axis, and the top element is the one with the lagging phase angle. Patterns are for 0° elevation angle over ideal ground. *(Courtesy* The ARRL Antenna Book.)**

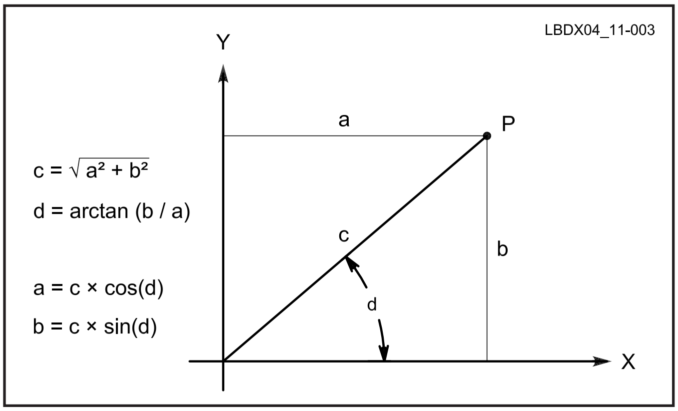



**Fig 11-3—Complex number representation.**

must always be represented as a point in a plane. A real number has one coordinate (the distance from the origin on the line) while the complex figure has two coordinates, which are necessary to unambiguously define its position in a plane.

- **Rectangular coordinates**: In a rectangular coordinate system, which in a plane consists of an X and a Y-axis, the X and the Y coordinates define the complex number. If the X-value is a and the Y-value equals b, the complex number is written as a + *j* b. The "j" indicates that the figure following is the Y-coordinate, which stands for the imaginary part.
- **Polar coordinates**: In a polar-coordinate system the position of the point representing the complex number is given by its distance to the coordinate origin and the angle of the vector going from the origin to the point, the angle with respect of the X-axis. The complex number in a polar coordinate system is written as c $\underline{/d°}$ where c = vector length and d = angle.

For some reason impedances are usually written in rectangular form as a + *j* b, while voltages and currents are most often represented in polar notation as c$\underline{/d°}$ . In the NEW LOW BAND SOFTWARE, complex values of Z, I and E are always expressed in both coordinate systems. With some simple trigonometry we can always convert from one system to another (see **Fig 11-3**, where conversion formulas are included).

## 3.3. Getting the Right Current Magnitude and Phase

There is a world of difference between designing an array on paper or with a computer modeling program and realizing it in real life. With single-element antennas (a single vertical, a dipole, etc) we do not have to bother about the feed current (magnitude and phase), as there is only one feed point anyway. With phased arrays things are vastly different.

First, we must decide which array to build. Once we do this, the problem will be how to achieve the right feed currents in all the elements (magnitude and phase angle). When we analyze an array with a modeling program, we notice that the feed-point impedances of the elements change from the value for a single element. If the feed-point impedance of a single quarter-wave vertical is 36 Ω over perfect ground, it is almost always different from that value in an array because of mutual coupling.

### 3.3.1. The effects of mutual coupling

Until about 10 to 15 years ago, few articles in Amateur-Radio publications addressed the problems associated with mutual coupling in designing a phased array and in making it work as it should. Gehrke, K2BT, wrote an outstanding series of articles on the design of phased arrays (Refs 921-925, 927). These are highly recommended for anyone who is considering putting up phased arrays of verticals. Another excellent article by Christman, K3LC (ex-KB8I) (Ref 929), covers the same subject. The subject has been very well covered in the 15th and later editions of *The ARRL Antenna Book*, where R. Lewallen, W7EL, wrote a comprehensive contribution on arrays. Today, Tom Rauch, W8JI, is a good teacher on principles and practical aspects of arrays in his excellent website (**www.w8ji.com/**), and his advice in these matters on the Topband reflector are much appreciated by all.

If we bring two (nearly) resonant circuits into the vicinity of each other, mutual coupling will occur. This is the reason that antennas with parasitic elements work as they do. Horizontally polarized antennas with parasitically excited elements are widely used on the higher bands. On the low bands the proximity of the ground limits the amount of control the designer has on the current in each of the elements. Arrays of vertical antennas, where each element is fed, overcome this limitation, and in principle the designer has an unlimited control over all the design parameters. With so-called *phased arrays*, all elements are individually and physically excited by applying power to the elements through individual feed lines. Each feed line supplies current of the correct magnitude and phase.

There is one frequently overlooked major problem with arrays. As we have made up our minds to feed all elements, we too often assume (incorrectly) there is no mutual coupling or that it is so small that we can ignore it. Taking mutual coupling into account complicates life, as we now have two sources of applied power to the elements of the array: parasitic coupling plus direct feeding.

#### *3.3.1.1. Self-impedance*

If a single quarter-wave vertical is erected, we know that the feed-point impedance will be 36 + *j* 0 Ω, assuming resonance, a perfect ground system and a reasonably thin conductor diameter. In the context of our array we will call this the *self-impedance* of the element.

#### *3.3.1.2. Coupled impedance*

If other elements are closely coupled to the original element, the impedance of the original element will change. Each of the other elements will couple energy into the original element and vice versa. This is often termed *mutual coupling* since each element affects the other. The coupled impedance is the impedance of an element being influenced by one other element and it is significantly different from the self-impedance in most cases.

#### *3.3.1.3. Mutual impedance*

The mutual impedance is a term that defines unambiguously the effect of mutual coupling between a set of two antenna elements. Mutual impedance is an impedance that cannot be measured. It can only be calculated. The calculated mutual impedances and driving impedances have been extensively covered by Gehrke, K2BT (Ref 923).

### *3.3.1.4. Drive impedance*

To design the correct feed system for an array, you must know the drive impedances of each of the elements, as well as the correct current magnitude and angle needed to feed the element(s).

### 3.3.2. Calculating the drive impedances

Mutual impedances are calculated from measured self-impedances and drive impedances. Here is an example: We are constructing an array with three λ/4 elements in a triangle, spaced λ/4 apart. We erect the three elements and install the final ground system, making the ground system as symmetrical as possible. Where the buried radials cross, we terminate them in a bus. Then the following steps are carried out:

1. Open-circuit elements 2 and 3. Opening an element will effectively isolate it from the other elements in the case of quarter-wave elements. (When using half-wave elements the elements must be grounded for maximum isolation and open-circuited for maximum coupling.)
2. Measure the self-impedance of element 1 (= Z11).
3. Ground element 2.
4. Measure the coupled impedance of element 1 with element 2 coupled (= Z1,2).
5. Open-circuit element 2.
6. Ground element 3.
7. Measure the coupled impedance of element 1 with element 3 coupled (= Z1,3).
8. Open-circuit element 3.
9. Open-circuit element 1.
10. Measure the self-impedance of element 2 (= Z22).
11. Ground element 3.
12. Measure the coupled impedance of element 2 with element 3 coupled (= Z2,3).
13. Open-circuit element 3.
14. Ground element 1.
15. Measure the coupled impedance of element 2 with element 1 coupled (= Z2,1).
16. Open-circuit element 1.
17. Open-circuit element 2.
18. Measure the self-impedance of element 3 (= Z33).
19. Ground element 2.
20. Measure the coupled impedance of element 3 with element 2 coupled (= Z3,2).
21. Open-circuit element 2.
22. Ground element 1.
23. Measure the coupled impedance of element 3 with element 1 coupled (= Z3,1).

This is the procedure for an array with 3 elements. The procedures for 2 and 4-element arrays can be derived from the above.

As you can see, measurement of coupling is done for pairs of elements. At step 15, you are measuring the effect of mutual coupling between elements 2 and 1, and it may be argued that this had already been done in step 4. It is useful, however, to make these measurements again to recheck the previous measurements and calculations. Calculated mutual couplings Z12 and Z21 (see below) using the Z1,2 and Z2,1 inputs should in theory be identical, and in practice should be within an ohm or so. The self-impedances and the driving impedances of the different elements should match closely if the array is to be made switchable.

The mutual impedances can be calculated as follows:

$$Z12 = \pm\sqrt{Z22 \times (Z11 - Z1{,}2)}$$

$$Z21 = \pm\sqrt{Z11 \times (Z22 - Z2{,}1)}$$

$$Z13 = \pm\sqrt{Z33 \times (Z11 - Z1{,}3)}$$

$$Z31 = \pm\sqrt{Z11 \times (Z33 - Z1{,}3)}$$

$$Z23 = \pm\sqrt{Z33 \times (Z22 - Z2{,}3)}$$

$$Z32 = \pm\sqrt{Z22 \times (Z33 - Z3{,}2)}$$

It is obvious that if Z11 = Z22 and Z1,2 = Z2,1, then Z12 = Z21. If the array is perfectly symmetrical (such as in a 2-element array or in a 3-element array with the elements in an equilateral triangle), all self-impedances will be identical (Z11 = Z22 = Z33), and all driving impedances as well (Z2,1 = Z1,2 = Z3,1 = Z1,3 = Z2,3 = Z3,2). Consequently, all mutual impedances will be identical as well (Z12 = Z21 = Z31 = Z13 = Z23 = Z32). In practice, the values of the mutual impedances will vary slightly, even when good care is taken to obtain maximum symmetry.

Because all impedances are complex values (having real and imaginary components), the mathematics involved are difficult. The MUTUAL IMPEDANCE AND DRIVING IMPEDANCE software module of the NEW LOW BAND SOFTWARE will do all the calculations in seconds. No need to bother with complex algebra. Just answer the questions on the screen.

**Fig 11-4** shows the mutual impedance to be expected for quarter-wave elements at spacings from 0 to 1.0 λ. The resistance and reactance values vary with element separation as a damped sine wave, starting at zero separation with both signs positive. At about 0.10 to 0.15-λ spacing, the reactance sign changes from + to –. This is important to know in order to assign the correct sign to the reactive value (obtained via a square root).

Gehrke, K2BT, emphasizes that the designer should actually measure the impedances and not take them from tables. Some methods of doing this are described in Ref 923. The published tables show ballpark figures, enabling you to verify the square-root sign of your calculated results.

After calculating the mutual impedances, the drive impedances can be calculated, taking into account the drive current (amplitude and phase). The driving-point impedances are given by:

$$Zn = \frac{I1}{In} \times Zn1 + \frac{I2}{In} \times Zn2 + \frac{I3}{In} \times Zn3 ... + \frac{In}{In} \times Znn$$

where n is the total number of elements. The number of equations is n. The above formula is for the nth element. Note

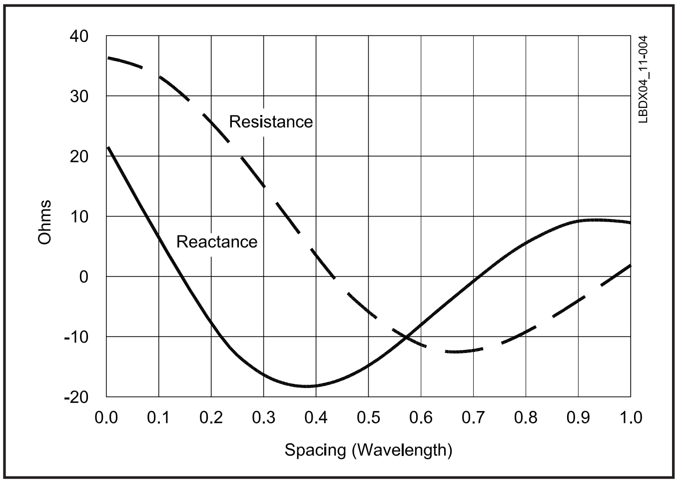



**Fig 11-4—Mutual impedance for two λ/4 elements. For shorter vertical elements (length between 0.1 λ and 0.25 λ), you can calculate the mutual impedance by multiplying the figures from the graph by the ratio $R_{rad}$/36.6 where $R_{rad}$ = the radiation resistance of the short vertical.**

also that Z12 = Z21 and Z13 = Z31, etc.

The above-mentioned program module performs the rather complex driving-point impedance calculations for arrays with up to 4 elements. The required inputs are:

1. The number of elements.
2. The driving current and phase for each element.
3. The mutual impedances for all element pairs.

The outputs are the driving-point impedances Z1 through Zn.

### *3.3.2.1. Design example*

Let us examine an array consisting of two λ/4 long verticals, spaced λ/4 apart and fed with equal magnitude currents, with the current in element 2 lagging the current in element 1 by 90°. This is the most common (though not necessarily the best) end-fire configuration with a cardioid pattern.

#### *Self impedance*

The quarter-wave long elements of such an array are assumed to have a self-impedance of 36.4 Ω over perfect ground. A nearly perfect ground system consists of at least 120 half-wave radials (see Chapter 9). For example, a system with only 60 radials may (depending on the ground quality) show a self-impedance on the order of 40 Ω.

#### *Coupled impedance*

We measured 37.5 + *j* 15.2 Ω.

#### *Mutual impedance*

The mutual impedances were calculated with the above-mentioned computer program: Z12 = Z21 = 19.76 – *j* 15.18 Ω. From the mutual impedance curves in Fig 11-4 it is clear that the minus sign is the correct sign for the reactive part of the impedance.

#### *Drive impedances*

The same software module calculates the drive impedances (also called feed-point impedances) of the two elements:

Z1= 55.8 + *j* 19.8 Ω for the – 90° element.
Z2 = 24.8 – *j* 19.8 Ω for the 0° element

We have now calculated the impedance of each element of the array, the array being fed with the current (magnitude and phase) as set out. We have used impedances that we have measured; we are not working with theoretical impedances.

The 2 EL AND 4 EL VERTICAL ARRAYS module of the NEW LOW BAND SOFTWARE is the perfect tool to guide you along the design of an array. You can enter your own values or just work your way through using a standard set of values.

### 3.3.3. Modeling the array

With the latest *NEC-3* or *NEC-4* based software, you can include a buried radial system, but for the design and evaluation of arrays, a *MININEC*-based modeling program, or even better a *NEC*-based program using a *MININEC*-type of ground (such as provided in *EZNEC*) will work, so long as we realize that we must add some equivalent series resistance to account for the ground-losses of the radial system. To simulate the

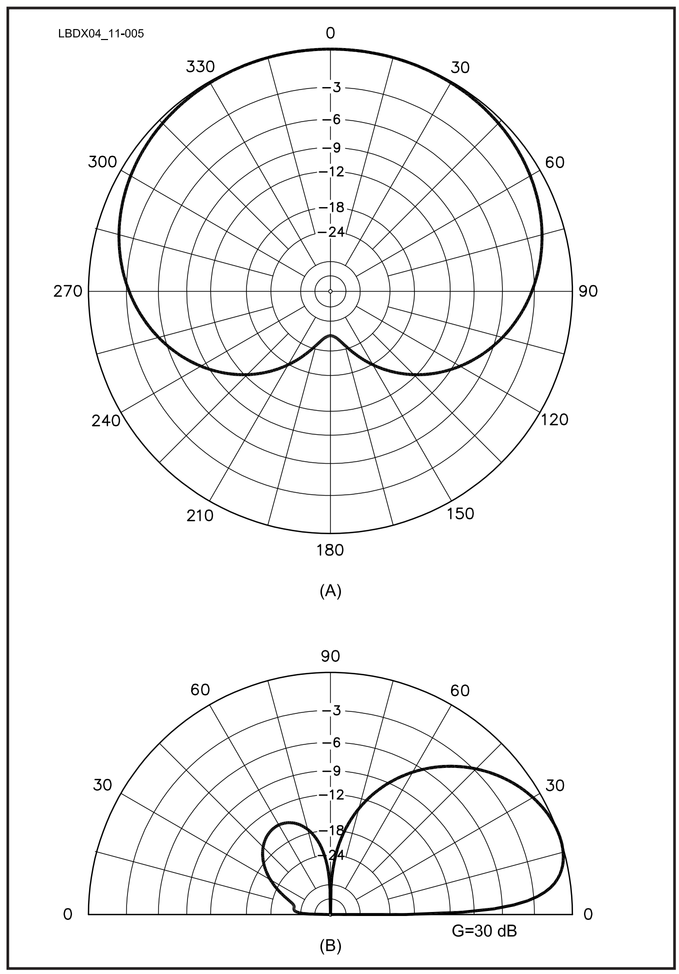



**Fig 11-5—Vertical and horizontal radiation patterns for the 2-element cardioid array, spaced 90° and fed with 90° phase difference. The pattern was calculated for very good ground with a radial system consisting of 120 radials, each 0.4 λ long (the equivalent ground resistance is 2 Ω). The gain is 3.0 dB compared to a single vertical over the same ground and radial system. The horizontal pattern at A is for an elevation angle of 19°.**

effect of a radial system consisting of 60 quarter-wave radials I inserted 4 Ω in series with the feed point of each antenna.

Modeling the cardioid antenna over *MININEC*-type ground with 4-Ω loss resistance included in each element, *EZNEC* comes up with the following impedances:

$Z1 = 55.0 + j\,22.7\ \Omega$
$Z2 = 26.5 - j\,19.5\ \Omega$

These are close to the values worked out with the NEW LOW BAND SOFTWARE, which were based on measured values of coupled and self impedances. The vertical and the horizontal radiation patterns for the 2-element cardioid array are shown in **Fig 11-5**.

## 3.4. Designing a Feed System

The challenge now is to design a feed system that will supply the right current to each of the array elements. As we now know the current requirements as well as the drive-impedance data for each element of the array, we have all the required inputs to design a feed system.

Each element will need to be supplied power through its own feed line. In a driven array each element either gets power, or it possibly delivers power into the feed system. During calculations we will sometimes encounter a negative feed-point impedance, which means the element is actually delivering power into the feed network. If the element impedance is zero, this means that the element can be shorted to ground. It then acts as a parasitic element.

Eventually all the feed lines will be connected to a common point, which will be the common feed point for the entire array. You can only connect feed lines in parallel if the voltages on the feed lines (at that point) are identical (in magnitude and phase)—the same as with ac power!

Designing a feed system consists of calculating the feed lines (impedance and length) as well as the component values of networks used in the feed system, so that the voltages at the input ends of the lines are identical. It is as simple as that.

The ARRL has published the original (1982) work by Lewallen, W7EL, in the last five editions of *The ARRL Antenna Book*. This material is a must for every potential array builder. However, there are other feed methods than the Lewallen method. Various feed systems are covered in the following sections of this book:

- Christman method
- Using flat lines
- Cross-fire principle
- Lewallen (quadrature fed arrays)
- Lewallen/Lahlum (any phase angle, any current ratio)
- Collins (hybrid coupler)
- Gehrke (broadcast approach)
- Lahlum/Gehrke (non current-forcing, L-network)

### 3.4.1. The wrong way

In just about all cases, the drive impedance of each element will be different from the characteristic impedance of the feed line. This means that there will be standing waves on the line. This has the following consequences:

- The impedance, voltage and current will be different in each point of the feed line.
- The current and voltage phase shift is not proportional to the feed line length, except for a few special cases (eg, a half-wave-long feed line).

This means that if we feed these elements with 50-Ω coaxial cable, we cannot simply use lengths of feed line as phasing lines by making the line length in degrees equal to the desired phase delay in degrees. In the past we have seen arrays where a 90° long coax line was inserted in one of the feed lines to an element to create a 90° antenna current phase shift. Let us take the example of the 2-element cardioid array (as described above) and see what happens (see **Fig 11-6**).

We run two 90° long coax cables to a common point. Using the COAX TRANSFORMER/SMITH CHART software module of the NEW LOW BAND SOFTWARE, we calculate the impedances at the end of those lines (I took RG-213 with 0.35 dB/100 feet attenuation at 3.5 MHz). Using round figures, the array element feed impedances, including 2 Ω of equivalent ground loss resistance, are:

$Z1 = 51 + j\,20\ \Omega$
$I1 = 1\ \text{A}\ \underline{/-90^\circ}$

From E = Z / I we can calculate (don't worry, the software does it for you):

$E1 = 54.8\ \underline{/-68.6^\circ}\ \text{V}$

and

$Z2 = 21 - j\,20\ \Omega$
$I2 = 1\ \underline{/0^\circ}\ \text{A}$
$E2 = 29\ \underline{/-43.6}\ \text{V}$

At the end of the 90° long RG-213 feed lines the impedances (and voltages) become:

$Z1' = 42.81 - j\,16.18\ \Omega$
$E1' = 50.89\ \underline{/0.39^\circ}\ \text{V}$
$I1' = 1.11\ \underline{/21.09^\circ}\ \text{A}$

and

$Z2' = 63.1 + j\,56.94\ \Omega$
$E2' = 50.37\ \underline{/89.61}\ \text{V}$
$I2' = 0.59\ \underline{/47.54}{}^\circ\text{A}$

If we make the line to the lagging element 180° long (90° plus the extra 90° to obtain an extra 90° phase shift), we end up with:

$Z1'' = 51.18 + j\,18.64\ \Omega$
$E1'' = 56.42\ \underline{/110.77}\ \text{V}$
$I1'' = 1.04\ \underline{/90.76}\ \text{A}$

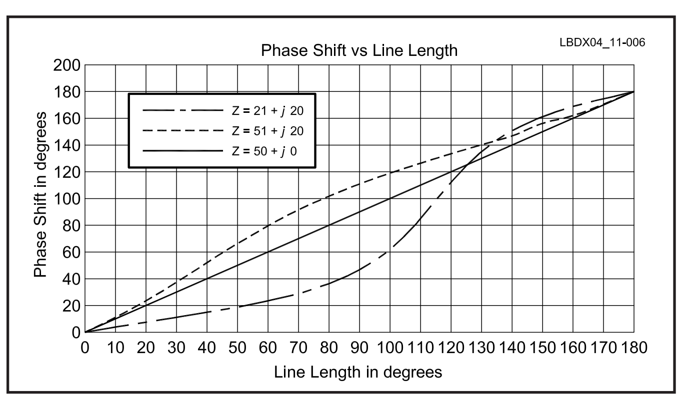



**Fig 11-6—Graph showing the current phase shift in a 50-Ω line (RG -213, on 80 meters), as a function of the load impedance. The loads shown are those for a 2-element cardioid array. Note that the phase shift does *not* equal line length, except when the line is terminated in its own characteristic impedance!**

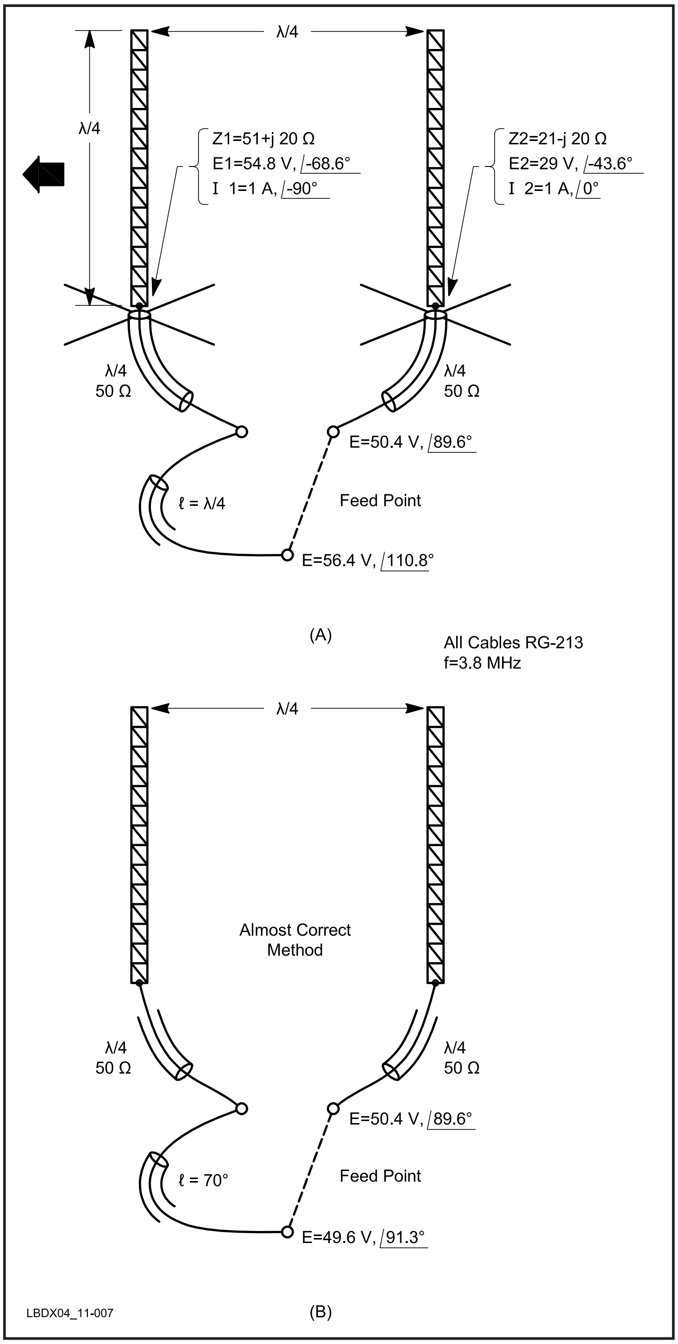



**Fig 11-7—At A, the incorrect way of feeding a 2-element cardioid array (90° phase, 90° spacing). Note that the voltages at the input ends of the two feed lines are not identical. In B we see the same system with a 70° long phasing line, which now produces almost correct voltages. The F/B ratio of existing installations will jump up by 10 or 15 dB, just by changing the line length from 90° to 70°.**

Note that E2′ and E1″ are not identical. This means we cannot connect the lines in parallel at those points without upsetting the antenna current (magnitude and phase). From the above voltages we see that the extra 90° line created an actual current phase difference of 90.76° – 21.09° = 68.67°, and *not* 90° as required.

The software module IMPEDANCES, CURRENTS AND VOLTAGES ALONG FEED LINES is ideally suited for analyzing this phenomenon. Look at the values of voltage and current as you scan along the line, and remember we want the right current phase shift and we want the same voltage where we connect the feed lines in parallel.

If you have such a feed system, do not despair. Simply by shortening the phasing line from 90° to 70°, you can obtain an almost perfect feed system. (See **Fig 11-7**.)

Watch out, if you want to use this system, make sure you have the same feed impedances as in the model above. How? By calculating the drive impedances as outlined in Section 3.3.2, or by carefully modeling your array, making sure you take into account all the small details!

### 3.4.2. Christman (K3LC) method

In the Christman, K3LC (ex-KB8I) method (Ref 929), we scan the feed lines to the different elements looking for points where the voltages are identical. If we find such points, we connect them together, and we are all done! It's really as simple as that. Whatever the length of the lines are, provided you have the right current magnitude and phase at the input ends of the lines, you can always connect two points with identical voltages in parallel. That's also where you feed the entire array.

Christman makes very clever use of the transformation characteristics of the feed lines. We know that on a feed line with SWR, voltage, current and impedance are different in every point of the line. The questions are now, "Are there points with identical voltage to be found on all of the feed lines?" and "Are the points located conveniently; in other words, are the feed lines long enough to be joined?" This has to be examined case by case.

It must be said that we cannot apply the Christman method in all cases. I have encountered situations where identical voltage points along the feed lines could not be found. The software module, IMPEDANCE, CURRENT AND VOLTAGE ALONG FEED LINES, which is part of the NEW LOW BAND SOFTWARE, can provide a printout of the voltages along the feed lines. The required inputs are:

- Feed-line impedance.
- Driving-point impedances (R and X).
- Current magnitude and phase.

Continuing with the above example of a 2-element configuration (90°spacing, 90° phase difference, equal currents, cardioid pattern), we find:

E1 = (155° from the antenna element) = 47.28 /86.1°V
E2 = (84° from the antenna element) = 47.27 /85.9°V

Notice on the printout that the voltages at the 180° point on line 1 and at the 90° point on line 2 are not identical (see Section 3.4.1), which means that if you connect the lines in parallel in those points, you will not have the proper current in the antennas.

We need now to connect the two feed lines together where the voltages are identical. If you want to make the array switchable, run two 84° long feed lines to a switch box, and insert a 155° – 84° = 71° long phasing line, which will give you the required 90° antenna-current phase shift. **Fig 11-8** shows the Christman feed method.

Of course the impedance at the junction of the two feed lines is not 50 Ω. Using the COAX TRANSFORMER/SMITH CHART software module, we calculate the impedances at the

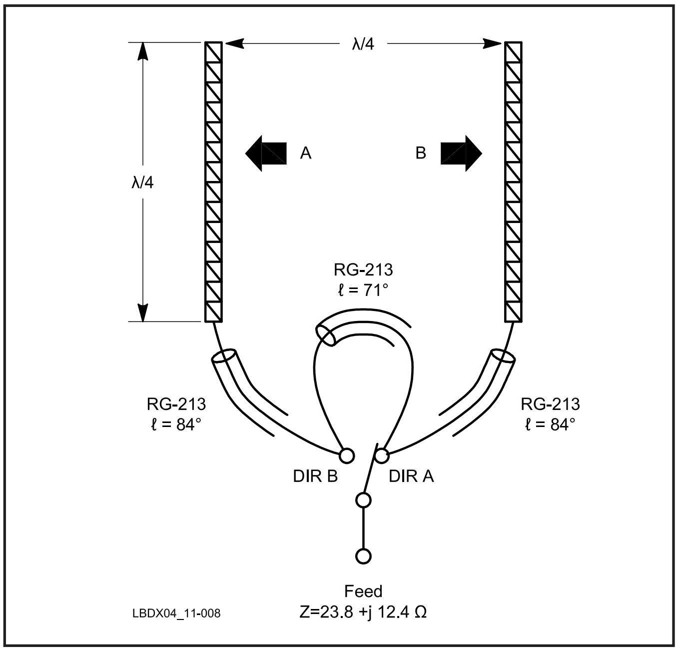



**Fig 11-8—Christman feed system for the 2-element λ/4-spaced cardioid array fed 90° out-of-phase. Note that the two feed lines are 84° long (not 90°), and that the "90° phasing line" is actually 71 electrical degrees in length. The impedance at the connection point of the two lines is 23.8 + *j* 12.4 Ω (representing an SWR of 2.3:1 for a 50-Ω line), so some form of matching network is desirable.**

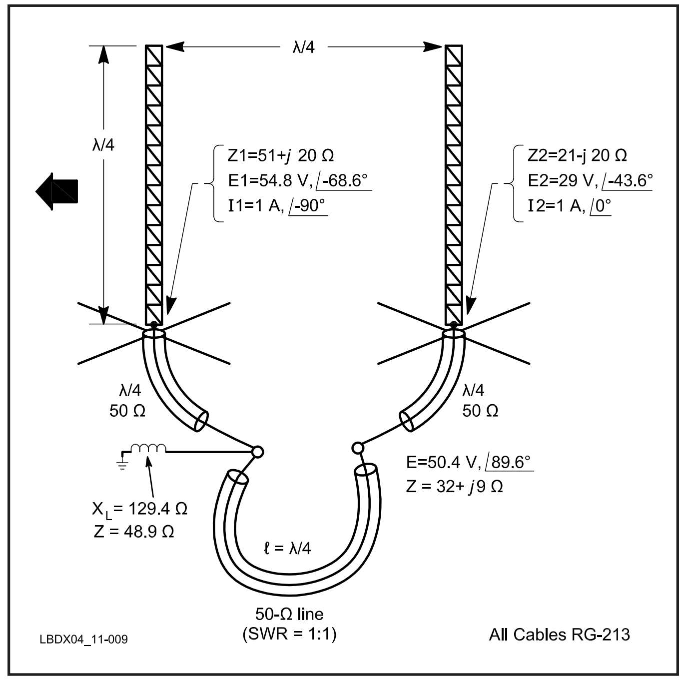



**Fig 11-9—Adding a shunt coil with a reactance of +129.4 Ω at the end of the λ/4 feed line going to the front element turns the impedance at that point into 48.9 Ω, very close to 50 Ω. Now we can insert a 50-Ω delay line and be assured that the phase shift equals the line length.**

input ends of the two lines we are connecting in parallel:

Z1end = 39 + *j* 12 Ω
Z2end = 50 + *j* 52 Ω

The software module PARALLEL IMPEDANCES calculates the parallel impedance as 23.8 + *j* 12.4 Ω. This is the feed-point impedance of the array. You can use an L network, or any other appropriate matching system to obtain a more convenient SWR on the 50-Ω feed line.

### 3.4.3. Using flat lines (SWR ~ 1:1) with "length = phase shift"

Let's go back to Fig 11-7. The impedance at the end of the quarter-wave line going to the front element is 42.81 – *j* 16.18 Ω. Maybe we can turn it in a purely resistive impedance of convenient value by connecting a reactance in parallel. Using the SHUNT/SERIES IMPEDANCE NETWORK module of the LOW BAND SOFTWARE, we can easily calculate the required parallel impedance to make it a purely resistive impedance. In this case it appears that putting an inductance of +129.4 Ω in parallel at that point, turns the impedance to 48.9 Ω, very close to 50 Ω. Let's do that, and now connect a quarter-wave phasing line from that point to the end of the quarter-wave line coming from the back element. As the line now operates with an SWR of very close to 1:1, phase difference equals line length, and we have exactly what we want.

**Fig 11-9** shows the layout of this system. If you want more phase shift, eg 120° to lift the notch off the ground (see Chapter 7) you simply make the phasing line 120° long. Note however that the element feed impedances shown are for 90° phase shift and that those are slightly different when you change the elevation angle.

It is obvious that such method can only be applied when you are lucky to find an impedance (after tuning out the reactance by a parallel element) that matches an existing feed-line impedance. You can, of course, use parallel feed line to obtain low impedances, and actually connect feed lines of different impedances in parallel (25 Ω = two 50-Ω lines in parallel; 30 Ω = a 50-Ω and a 75-Ω line in parallel; 37.7 Ω = two 75-Ω in parallel).

### 3.4.4. The Cross Fire (W8JI) principle

In a "standard" array, for example as shown in Section 3.4.2 and 3.4.3, the feed line goes to the back element, and the front element is fed via a phasing line. Let us analyze what happens in such a design when we change frequency away from the nominal design frequency. Assume we have a 2-element end-fire array, spaced exactly λ/4 (90°) and with exactly 90° phase shift (this is by far not the best arrangement!). Our notch elevation angle will be 0° (see Chapter 7). If we increase the frequency by 5%, the spacing becomes 94.9° and the phasing becomes also larger (if the lines are relatively flat also about 5% longer). But, in order to maintain the zero notch angle at ground level, we need the phasing line to be *shorter* by about 5%. This mechanism limits the usable bandwidth in such arrays. In simple 2-element arrays this usually is not a problem, but in more complex arrays using four or more elements it can become a key design factor.

Tom Rauch, W8JI, pointed out that we can also use the cross-fire principle feed method, where we feed the array at the front element using a phase inverter (a 180° transformer) and feed the back element with a phasing line that is complementary in length to the required phasing angle (see Chap-

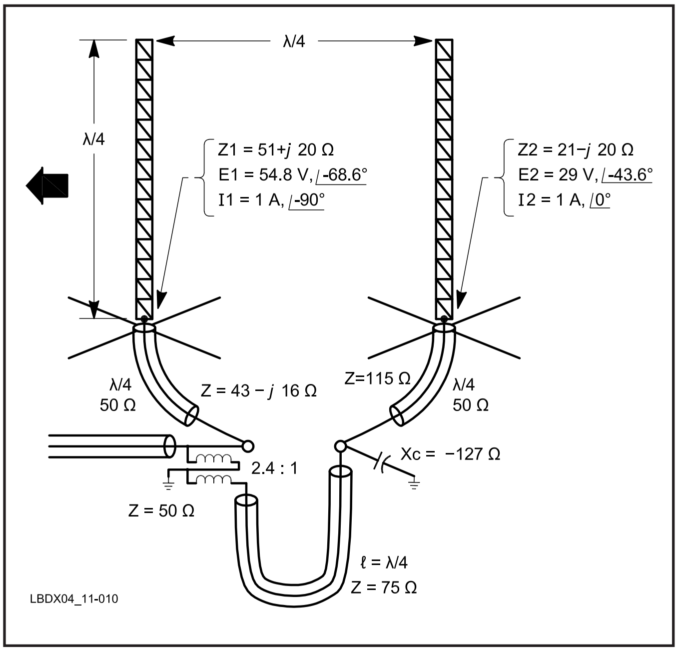



**Fig 11-10—While all other feed methods feed the back element directly and provide phase delay via coaxial cable or a network to the front element, the cross-fire feeding system does the opposite. It makes uses of a 180° phase-inverter transformer to achieve a feed system that guarantees that the phase delay remain correct when the frequency is changed. See text for details.**

ter 7). In this case the phase-shift transformer produces a 180° shift over a wide frequency range. At a frequency that is 5% higher than the design frequency, the phase shift produced by the phasing line becomes about 95° long. Subtracting this value from the 180° phase shift obtained by the transformer, the phase difference becomes 85° at the higher frequency. With this cross-fire principle the tracking is achieved, which is exactly what we want.

In **Fig 11-10** we see that we will have to put the phasing line in the feed line going to the back-element. The impedance at the end of the quarter wave line to the elements is 63.1 + *j* 56.94 Ω. Using the SHUNT/PARALLEL IMPEDANCE section from the NEW LOW BAND SOFTWARE program, we find that a parallel capacitor with an impedance of −127 Ω will turn the impedance into 115 Ω, not exactly a common coaxial cable impedance. But what if we used a quarter-wave 75-Ω feed line for achieving a 90° phase shift? This will work but because the antenna impedance is not the same as the load impedance, the typical quarter-wave impedance transformation will occur. The impedance at the end of the line will be (75 × 75)/115 = 49 Ω. This means that there will be a voltage transformation of 115/49 = 2.3:1. In this particular setup, we will need to use a 180°-phase-shift transformer that has a transformation ratio (turns ratio) of 2.3:1 if we want to end up with equal current magnitudes at both elements.

Would you ever want to go through this procedure to achieve tracking? No, because tracking is limited anyhow by the variation in element feed impedances as you change frequency. This principle holds very well, however, when you are using elements that show little or no change in feed impedance when the frequency is changed, which is what occurs with many receiving antennas, as explained in Chapter 7.

This principle can also be used with complex arrays (4 elements and more) to achieve better bandwidth. Such designs are far from being "plug and play" and are explained for the reader to understand the principle rather than to serve as a building kit! For an application of this principle see Section 4.7.2.

### 3.4.5. Using an L network to obtain a desired shift

#### *3.4.5.1. Current Forcing:*

Roy Lewallen, W7EL, uses a method that takes advantage of the specific properties of quarter-wave feed lines (Lewallen calls it *current-forcing*). This method is covered in great detail by W7EL in recent editions of *The ARRL Antenna Book*.

A quarter-wave feed line has the following wonderful property, which is put to work with this particular feed method: The magnitude of the input current of a λ/4 transmission line is equal to the output voltage divided by the characteristic impedance of the line. It is independent of the load impedance. In addition, the input current lags the output voltage by 90° and is also independent of the load impedance.

#### *3.4.5.2. Using a simple L-network to obtain the right phase shift*

The method of using an L-network to obtain the proper phase shift was also introduced by Lewallen, W7EL. The original Lewallen feed method could only be applied to antennas fed in quadrature, which means antennas where the elements are fed with phase differences that are a multiple of 90°. Later the L-network technique approach was made more flexible, and the equations were made available where you could calculated the L-network for arrays where the L-network feeds more than one element, as well as for arrays where the current magnitude is not the same in all elements. Robye Lahlum, W1MK, worked out the following equations for such an L-network, including any arbitrarily chosen phase angle (no longer only multiples of 90°).

Let's have a look at **Fig 11-11.** This is an example of a

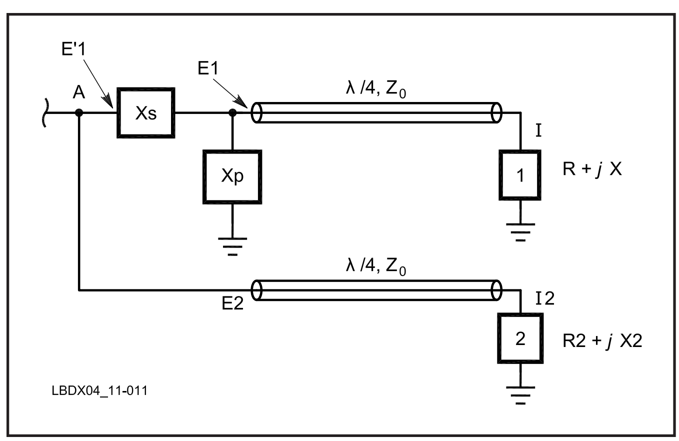



**Fig 11-11—Basic layout of the L-network phasing system developed by R. Lewallen, W7EL, and enhanced for any phase angle by R. Lahlum, W1MK.**

2-element array, where element 2 is fed directly, and element 1 is fed through an L-network. Both elements are fed through quarter-wave current-forcing feed lines—although this is not strictly necessary as explained in Section 3.4.9—but it makes measuring and tuning easier.

Voltage E1, at the end of the feed line going to element 1 is transformed in the L-network to E1′. The transformation is:

$$E1' \rightarrow k \times E1 \underline{/\theta°} \quad \text{(Eq 11-1)}$$

The k factor is related to the transformation's magnitude and the desired phase shift is represented by the angle θ. Obviously, we want to connect the input of the L-network (where the voltage is E1′) to the input of the quarter-wave feed line going to element 2, where the voltage is E2.

We can connect those two points together, if the voltages in those points are identical. In other words if:

$$E2 = k \times E1 \underline{/\theta°} \quad \text{(Eq 11-2)}$$

The condition for this to apply is:

$$X_S = \frac{-\sin\theta \times {Z_0}^2}{n \times k \times R} \quad \text{(Eq 11-3)}$$

$$X_P = \frac{Xs}{\left[\frac{n \times X \times X_s}{{Z_0}^2} - 1 + \frac{\cos\theta}{k}\right]} \quad \text{(Eq 11-4)}$$

Theta (θ) is the desired difference between the current phase angle at the element fed through the L-network and the phase angle at the input of the network. The phase angle is responsible for a time delay, and q must be negative. If necessary subtract 360° to obtain a negative value. Make sure you do not invert signs! Follow the examples given to understand the procedure.

The letter k is the ratio of the current supplied to the element in the branch fed through the L-network (in this case it is feed current magnitude of element 1), versus the current in the element fed directly (in this case, element 2).

The letter n is the number of identical elements (with identical feed currents) that are fed through the branch containing the L-network (see **Fig 11-12**) for a case where two elements are fed with identical current magnitudes and phases.

$Z_0$ is the characteristic impedance of the quarter-wave (or 3λ/4 or 5λ/4, etc) current-forcing feed lines.

R is the real part of the feed-point impedance of one of the identical element(s).

X is the imaginary part of the feed-point impedance (Z = X + *j* X).

$X_S$ is the impedance of the series element in the L-network.

$X_P$ is the impedance of the parallel (shunt) element in the L-network.

These apply under all circumstances where you feed the elements via current-forcing feed lines. The impedance R + *j* X is *not* the impedance at the end of the feed line but the feed-point impedance of an antenna element.

The equations do not work for 0° or 180°, but for 0° you do not need a phase-shifter and for 180° we have the choice between a half-wave long feed line or a 180°-phase-reversal transformer (see Section 4.15.7).

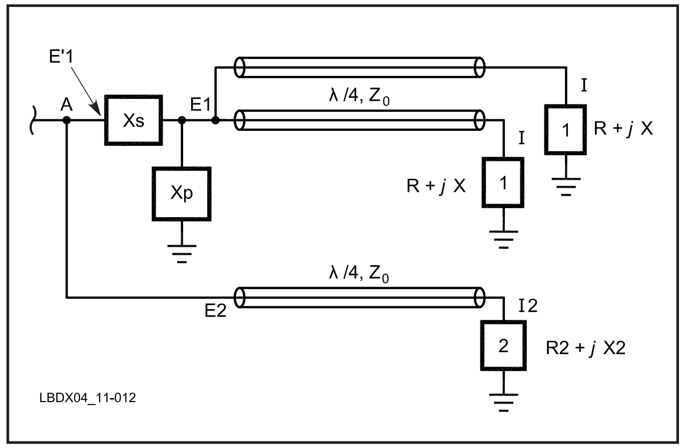



**Fig 11-12—In this particular case the L-networks feeds two elements with identical feed currents. All you need to do is enter n = 2 in the *Lahlum.xls* spreadsheet.**

Note that in these equations no consideration was given to the losses in the feed lines nor in the network. Under most real-life conditions these losses are small on the low bands. We can, however, do the calculation including cable losses as well (see Sections 3.4.5.4).

Also assuming no feed-line losses, we can easily calculate the input impedance at the input side of the L-network. The parallel input impedance components at the input of the network are:

$$R_{par} = \frac{{Z_0}^2}{k^2 \times n \times R} \quad \text{(Eq 11-5)}$$

$$X_{par} = \frac{X_S}{1 - k \times \cos\theta} \quad \text{(Eq 11-6)}$$

These values must be converted to their equivalent series-input impedances:

$$R_{ser} = \frac{R_{par} \times {X_{par}}^2}{{R_{par}}^2 + {X_{par}}^2} \quad \text{(Eq 11-7)}$$

and

$$X_{ser} = \frac{{R_{par}}^2 \times X_{par}}{{R_{par}}^2 + {X_{par}}^2} \quad \text{(Eq 11-8)}$$

This parallel-to-series calculation can also be done using the using the RC/RL transformation module of the NEW LOW BAND SOFTWARE program.

### *3.4.5.3 The Lahlum.xls spreadsheet tool*

I wrote an *Excel* spreadsheet (*Lahlum.xls*) that is on the CD-ROM bundled with this book. This tool allows you to calculate the values of the L network, as well as the resulting input impedance of the branch with the L-network. Usage is simple and self-explanatory. The spreadsheet uses the formulas shown in Section 3.4.5.2.

### *3.4.5.4. Two-element end-fire array in quadrature feed*

In this example in **Fig 11-13** for a 2-element end-fire array from Section 3.4.1, the L-network goes to one element (in a Four-Square it may drive two elements), so enter 1 for the number of elements. $Z_0$ is the characteristic impedance of the quarter-wave line going from the L-network to the element(s). R and X are the real and the imaginary values of the impedance of the element at the end of that line (in our case R = 51 and X = +20). For the moment enter k = 1, meaning that the current magnitude in the elements is identical. Use theta = (–90) – (0) = –90°.

As explained above, the formulas used in the spreadsheet assume no cable loss. If you want to calculate the L-network values and include cable loss, you can calculate the impedance at the end of the current-forcing feed line, using the COAX TRANSFORMER/SMITH CHART module of the NEW LOW BAND SOFTWARE, and use the option "with cable losses." You can also use a transmission-line program such as ARRL's *TLW*. Once you know the impedance at the end of the feed lines, you can calculate the L-network component values using the second part of the spreadsheet (called: "For system NOT USING current-forcing, or if using "real" quarter-wave lines").

The first part of the *Lahlum.xls* spreadsheet calculates without taking into account cable losses. **Fig 11-14** shows the feed network for the case where cable losses are included. Note that the difference in L-network values is very small. In most cases the lossless calculation will suffice. In most of the examples in this chapter, thus, we will use lossless calculations.

For a 2-element end-fire array we normally feed the back element directly, with the exception of feeding using the cross-fire principle (see Section 3.4.4.). We can, however, feed the front element directly and the back element with a phase shift. In the case of quadrature feeding, this is +90°, which equals +90 – 360 = –270°. We can achieve the –270° phase shift by designing an L-network to do just that, or we can do this using an L-network that takes care of –90°, followed by a half-wave of feed line, for another 180°. When *Lahlum.xls* is used with θ = 270°, the resulting L-network values are 352.0 pF and 104.7 µH. The inductance required is rather high, which is not desirable. If however we replace –270° with –90°, and add a half-wave feed line at the input of the L-network, we end up with much more attractive component values of 687.2 pF and 5.0 µH.

In many of the phased arrays described in this chapter, the rear element has a very low feed impedance, often with a negative value for the series resistance. At the end of the λ/4 current-forcing feed line, the impedance becomes very high. If we design a feed system that includes an L-network in this branch, we will very often end up with extreme component values. If the reactances are very high, the Q will be high and bandwidth very low. In many cases we will see reactances change from high positive values to high negative values with just a small change in frequency. This situation must be

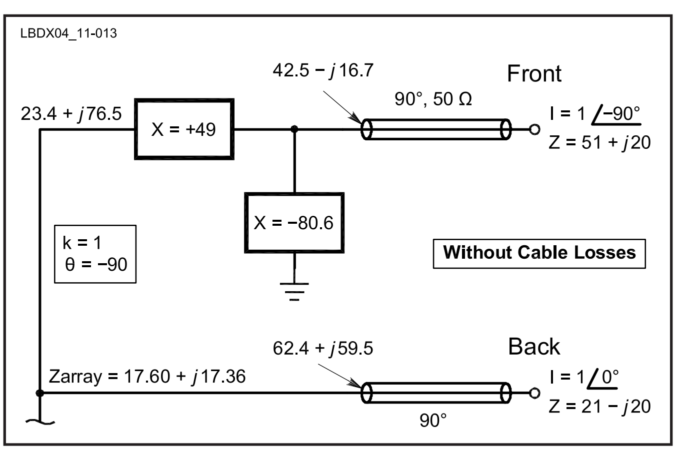



**Fig 11-13—Lewallen/Lahlum feed system for the 2-element cardioid array from Fig 11-10. Calculations were done assuming zero cable losses.**

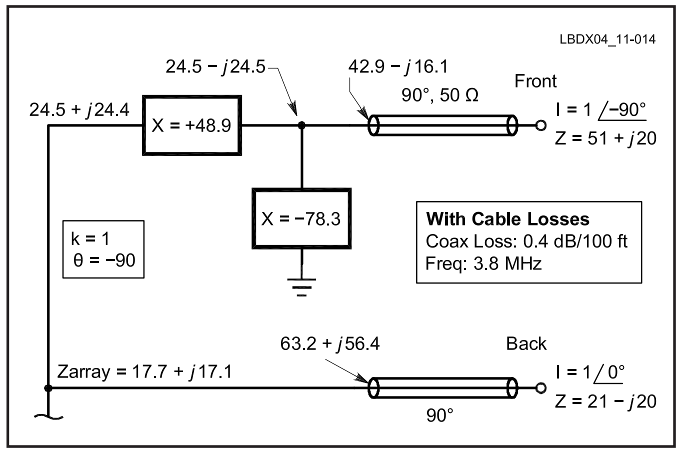



**Fig 11-14—Lewallen/Lahlum feed system for the 2-element cardioid array from Fig 11-10. Calculations were done including cable losses. Note the minute difference between the lossless and the "real-world" calculation results in Fig 11-13.**

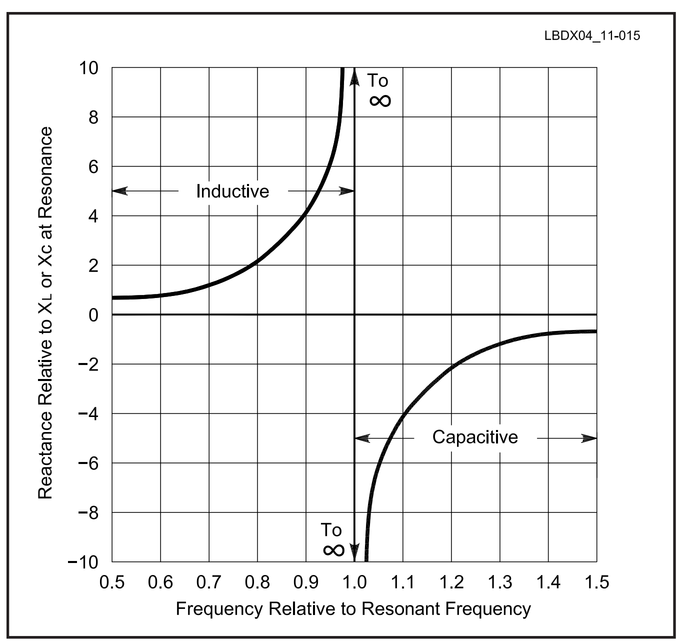



**Fig 11-15—This demonstrates that the change in reactance is much greater near resonance than far away.**

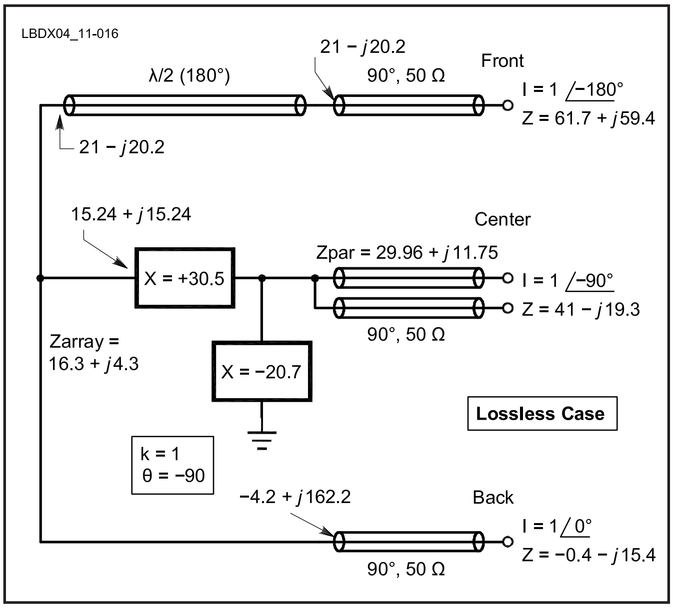



**Fig 11-16—Feed system for the quadrature-fed Four-Square, using 50-Ω current-forcing feed lines.**

avoided. Therefore it is always best to feed the rear element directly, and the center and front elements via L-networks.

**Fig 11-15** shows how the reactance near resonance abruptly changes from inductive to capacitive, and also demonstrates that the relative change in reactance is much greater in that area than farther away. It's a good rule of thumb to design a network where the absolute value of the component reactances are not larger than about 250 Ω.

It's a good idea always to work out all the alternative feed systems. You can do these exercises with 50 and with 75-Ω cable. And if there are phasing angles involved that are larger than 180°, you can use a half-wave coax cable to the 180° part (see **Fig 11-16**). For each alternative, look at the total array feed impedance and at the L-network component values.

### *3.4.5.5. Using a different $Z_0$ (current-forcing feed-line impedance)*

The example of Figs 11-13 and 11-14 results in a relatively low array feed impedance (~ 17.6 + *j* 17.4 Ω). We could do the same exercise using 75-Ω feed lines, and we will see a higher feed impedance. How much higher? Robye, W1MK, pointed out a simple rule-of-thumb (not 100% correct but a close indicator):

Z(feed–75 Ω) = Z(feed–50 Ω) × [75/50] × 2 = Z(feed–50 Ω) × 2.25 (Eq 11-9)

In our example the estimated (lossless) impedance, according to this rule is 2.25 × (17.6 + *j* 17.4) = 39.6 + *j* 39.15 Ω. If we do the detailed calculations, the feed impedance, using 75-Ω element feed lines, turns out to be: 39.6 + *j* 39.1 Ω, which confirms the simple rule above.

In this particular case it would certainly be better using 75-Ω element feed lines. Note that in many arrays using 75-Ω feed lines achieves an overall network drive impedance closer to 50 Ω than is the case when using 50-Ω lines.

The 39.6 + *j* 39.1 Ω can be matched pretty well to either a 50-Ω or a 75-Ω feed line to the shack using a series capacitor with X = –39.1 Ω, the feed impedance becomes 39.1 Ω, which results in an acceptable 1.25:1 SWR for 50-Ω coax. Using a parallel capacitor with a reactance of –78 Ω, results in a feed impedance of 78 Ω, giving a good match to a 75-Ω feed line, if you'd like to use that.

### *3.4.5.5. Calculation of array feed impedance*

There are two ways of doing this: without losses and with losses. In most cases the lossless way will suffice, but I will explain both ways.

### *3.4.5.6. Without losses: Using the Lahlum.xls spreadsheet:*

See Section 3.4.5.2 for the formulas, but the top part of the spreadsheet does all the work. Let's do it, step by step:

$R_{par} = Z_0^2/(n \times R \times k^2)$

and

$X_{par} = X_S/(1 - k \times \cos\theta)$

In this case k = 1 and θ = –90° so the formula becomes:

$X_{par} = Z_0^2/R$

and

$X_{par} = X_S$

Using the figures from the above example we have:

$R_{par} = (50 \times 50)/51 = 49\ \Omega$

and $X_{par} = 49\ \Omega$

These values must be converted to their series-equivalent input impedances using the following formulas:

$R_{ser} = (R_{par} \times X_{par}^2)/(R_{par}^2 + X_{par}^2) = (49 \times 49 \times 49)/(49 \times 49 + 49 \times 49) = 24.5\ \Omega$

$X_{ser} = (R_{par}^2 \times X_{par})/(R_{par}^2 + X_{par}^2) = (49 \times 49 \times 49)/(49 \times 49 + 49 \times 49) = 24.5\ \Omega$

The transformation from parallel to serial impedance (and vice versa) can also be calculated using the RC/RL transformation module of the LOW BAND SOFTWARE program. Now we connect this impedance in parallel with 62.4 + *j* 59.5 Ω. The result is 17.60 + *j* 17.36 Ω.

### *3.4.5.7. Including losses:*

Z1 = 51 + *j* 20 Ω
Z2 = 21 – *j* 20 Ω

Using the COAX TRANSFORMER/SMITH CHART module of the NEW LOW BAND SOFTWARE, we calculate the transformed impedances at the end of 90° long feed lines (Vf = 0.66, attenuation = 0.3 dB/100 feet, at F = 3.8 MHz):

Z1' = 42.95 – *j* 16.1 Ω
Z2' = 63.2 + *j* 56.4 Ω

Now – *j* 80.6 Ω in parallel with 42.5 – *j* 16.7 Ω = 24.49 – *j* 24.53 Ω. This is in series with + *j* 49 Ω, yielding 24.49 + *j* 24.53 Ω. Now, we connect this impedance in parallel with 63.2 + *j* 56.4 Ω and the result is 17.67 + *j* 17.12 Ω.

This calculation includes cable losses but not the losses from the L-network components. Note that this values is very close to what we calculated in the lossless case.

### *3.4.5.8. Tutorial:*

The 2 EL AND 4 EL VERTICAL ARRAYS module of the NEW LOW BAND SOFTWARE is a tutorial and engineering program that takes you step by step through the design of a 2-element cardioid type phased array (and also the famous Four-Square array, which I'll describe later in this chapter). The results as displayed in that program will be slightly different from the results shown here, since the software uses lossless feed lines.

### 3.4.5.9. The quadrature-fed Four-Square

Let's assume we have obtained the following feed impedance values through modeling a Four-Square array:

Z1 = 61.7 + *j* 59.4 Ω (at the front element, fed with a −180° current phase angle)

Z2 = Z3 = 41 – *j* 19.3 Ω (the center elements, both fed with a −90° current phase angle)

Z4 = –0.4 – *j* 15.4 Ω (at the 0° element, the back element)

Note that the –0.4-Ω resistive part of the feed impedance Z4 means that the antenna is not *taking* power from the feed network, bur rather *delivering* power to it. This is excess power due to mutual coupling to the other elements. Note also that in a lossless calculation such a negative (usually very low) value will show up as a negative (high) value at the end of the λ/4 feed line. If, however, the nominal value is low, and the cable attenuation is taken into consideration, a small negative R-value at the antenna can turn up a high positive R-value at the other end. This is due to the effect of cable loss.

Note also that in this array, as is the case in most multi-element arrays, the SWR of the feed line going to the back element is *very* high, which normally causes a lot of additional power loss due to SWR. But in this case, the power flow is so small into the feed line to the back element that it does not matter much. High SWR, but no power flow, results in very little watts being lost. If you look at the resistive part of the equivalent-parallel resistance (several thousand Ω) at the end of the λ/4 feed line, any reduction in the exact value due to losses would cause very little increase in input power to get the same current to flow into the loads. This means that you can use the lossless model to calculate the feed system.

As explained for the 2-element end-fire array we can design the feed system in different ways. The most common approaches are:

- Feeding the back element directly, the front element via a 180° phase shift line (λ/2) and the central elements via a L-network "from the back element," all of this with 50-Ω, λ/4 feed lines. See **Fig 11-16**.
- Identical as above, but with 75-Ω feed lines. See **Fig 11-17**.

There is no absolute need to feed the back element directly and the middle and front via a phasing system. You could feed the center elements directly and the front with a −90° phasing system (L network) and the back element with a −270° phasing system. In a third alternative you could feed the front directly, the center elements with –270° phase shift and the back with −180° phase shift.

Each solution will have different L-network component values and a different array input impedance and different values for the L-network components. You can then select the network with the most manageable network component values and the most attractive feed impedance (avoid values below 10 Ω).

It's a good idea always to work out all the alternative feed systems. In the case of a Four-Square array you can use either

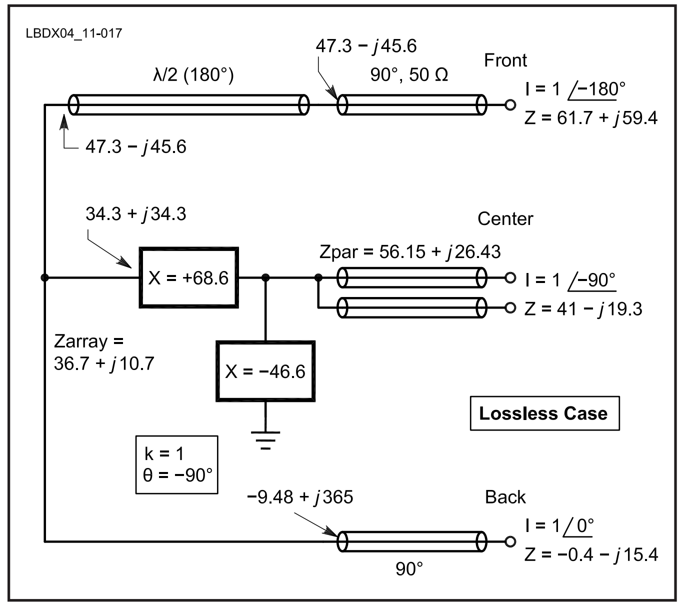



**Fig 11-17—Same feed system but using 75-Ω feed lines. This results in a significantly higher feed impedance, which is desirable.**

the branch to the back element as the *reference* branch, which is fed directly, or the branch to the center element or even the branch to the front element. You can do these exercises with 50-Ω and with 75-Ω cable. And if there are phasing values involved that are larger than 180°, you can use a half-wave coax cable to the 180° part (see Fig 11-16).

**Table 11-1** and **Table 11-2** show the *Lahlum.xls* results for a 75-Ω and for a 50-Ω system impedance, if we apply R = 41, X = –19.3, n = 2 and λ = –90°. It is obvious that the 75-Ω solution is the better one, since it results in a much more convenient array feed impedance.

### 3.4.5.10. The example of a three-in-line end-fire array with binomial current distribution.

You can also use the Lewallen feed system in arrays using different current magnitudes on each of the elements. Using parallel cables is not the right solution with the Lewallen method. The formulas, as given above, assume that the quarter-wave feed lines to all the elements in the array have the same impedance. If the current magnitude of the element fed through the L network needs to be different from the magnitude of the current to the other elements in the array, the appropriate current can be achieved by specifying the correct k-value in the *Lahlum.xls* spreadsheet or in the formulas from Section 3.4.5.

Let's work out the example of the 3 elements in-line array, each spaced λ/4, fed in 90° increments, but with the center element fed with double the current magnitude. See **Fig 11-18**.

Front element: Z1 = 76.1 + *j* 51 Ω
Center element: Z2 = 26.3 – *j* 0.4 Ω
Back element: Z3 = 15 – *j* 22.6 Ω

In the *Lahum.xls* spreadsheet in **Table 11-3** we enter k = 2, which means that the element(s) fed through the L-networks will have twice the current magnitude as the

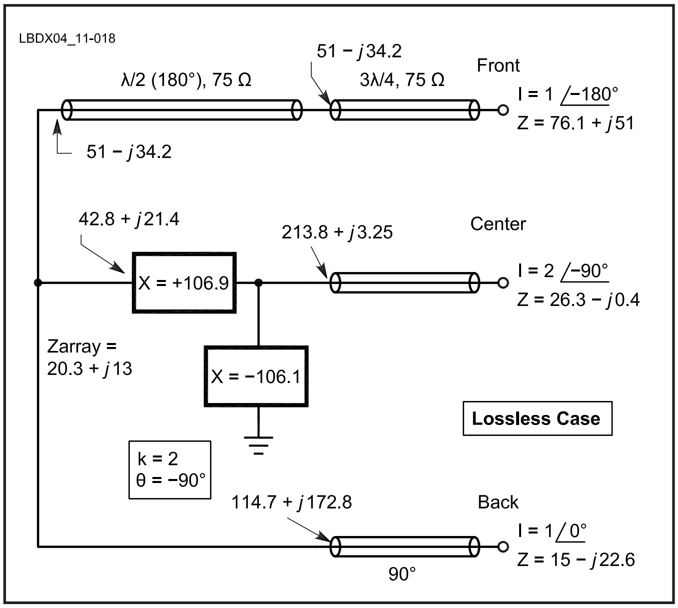



**Fig 11-18—Classic configuration with direct feed to the back element. Specifying k = 2 in the *Lahlum.xls* spreadsheet allows us to double the feed current magnitude without having to resort to paralleled feed lines.**

reference element in the array.

Fig 11-18 uses 75-Ω feed lines and results in an array feed impedance of 20.3 + *j* 13 Ω. Using 50-Ω feed lines, the array impedance would be approx 2.25 times lower, which is certainly not the best solution! Hence 75 Ω is recommended.

If we had included losses, the real part of the feed impedance would have been slightly higher, since you need more driving power into the feed system to get the same amount of radiated power.

### *3.4.5.10.1 Calculating the array input impedance*

In order to prove that the real part of the input impedance would indeed be higher, we will carry out a calculation in a real-world environment. Note that in order to reach the center of the array (which is necessary if you want to switch directions) you will need 3λ/4 feed lines, as the element physical spacing is λ/4.

Let's do some impedance calculations using the applicable modules of the NEW LOW BAND SOFTWARE. Using the COAX TRANSFORMER/SMITH CHART module we first calculate the impedances at the end of our 3λ/4 feed lines (5λ/4 feed line to front element). As explained earlier, a lossless calculation will do

Front element: Z1 = 76.1 + *j* 51 Ω → 79.1 + *j* 27.3 Ω
Center element: Z2 = 26.3 – *j* 0.4 Ω → 180.7 + *j* 2.2 Ω
Back element: Z3 = 15 – *j* 22.6 Ω → 128.9 + *j* 134.9 Ω

I used 75-Ω coax (Vf = 0.8) with a loss of 0.2 dB/100 feet for the calculation (design frequency = 1.8 MHz). Next we calculate the parallel impedance caused by the parallel reactance Xp1, using the module PARALLEL IMPEDANCES (T-JUNCTION).

Xp1 calculates as – *j* 106 Ω in parallel with 180.7 + *j* 2.2 Ω, which gives 46.8 – *j* 79.2 Ω. Adding + *j* 106.9 Ω in series yields: 46.8 + *j* 27.78 Ω.

**Table 11-1**
**Lahlum.xls spreadsheet results (see Fig 11-17):**

| | |
|---|---|
| *INPUT DATA* | |
| n elem | 2.00 |
| Z | 75.00 ohm |
| $R^0$ | 41.00 ohm |
| X | –19.30 ohm |
| k | 1.00 |
| theta | –90.00 deg |
| freq | 3.80 MHz |
| *RESULTS* | |
| Xs | 68.60 ohm |
| Xp | –46.64 ohm |
| Series elem | 2.9 µH |
| Par elem | 898.4 pF |
| Rpar | 68.60 ohm |
| Xpar | 68.60 ohm |
| Rser | 34.30 ohm |
| Xser | 34.30 ohm |

**Table 11-2**
**Lahlum.xls spreadsheet results (see Fig 11-16):**

| | |
|---|---|
| *INPUT DATA* | |
| n elem | 2.00 |
| Zo | 50.00 ohm |
| R | 41.00 ohm |
| X | –19.30 ohm |
| k | 1.00 |
| theta | –90.00 deg |
| freq | 3.80 MHz |
| *RESULTS* | |
| Xs | 30.49 ohm |
| Xp | –20.73 ohm |
| Series elem | 1.3 µH |
| Par elem | 2021.4 pF |
| Rpar | 30.49 ohm |
| Xpar | 30.49 ohm |
| Rser | 15.24 ohm |
| Xser | 15.24 ohm |

**Table 11-3**
**Lahlum.xls spreadsheet results (see Fig 11-18):**

| | |
|---|---|
| *INPUT DATA* | |
| n elem | 1.00 |
| Zo | 75.00 ohm |
| R | 26.30 ohm |
| X | –0.40 ohm |
| k | 2.00 |
| theta | –90.00 deg |
| freq | 3.80 MHz |
| *RESULTS* | |
| Xs | 106.94 ohm |
| Xp | –106.13 ohm |
| Series elem | 4.5 µH |
| Par elem | 394.8 pF |
| Rpar | 53.47 ohm |
| Xpar | 106.94 ohm |
| Rser | 42.78 ohm |
| Xser | 21.39 ohm |

For the back element we have an impedance of 128.9 + *j* 134.91 Ω at the end of the 3λ/4 feed line. For the front element we have an impedance of 79.1 + *j* 37.3 Ω at the end of the 5λ/8 feed line.

All three are in parallel, so Ztot = 24.5 + *j* 14.9 Ω**.** This is what we expected. Compared to the value we calculated without losses (20.3 + *j* 13) the feed impedance goes a little higher when losses are included.

### 3.4.5.11. Input impedance at the input side of the L-network

If we neglect the effect of losses in the feed lines, we can also calculate the input impedance using the $R_{ser}$ and $X_{ser}$ values from the *Lahlum.xls* worksheet. $R_{ser}$ = 42.78 Ω and $X_{ser}$ = 21.39 Ω

These values are somewhat lower than those calculated considering cable losses (46.8 + *j* 27.78 Ω) The difference is relatively high because in this case we are using 270° feed lines, which represents more loss. But for all practical purposes the lossless calculations are adequate.

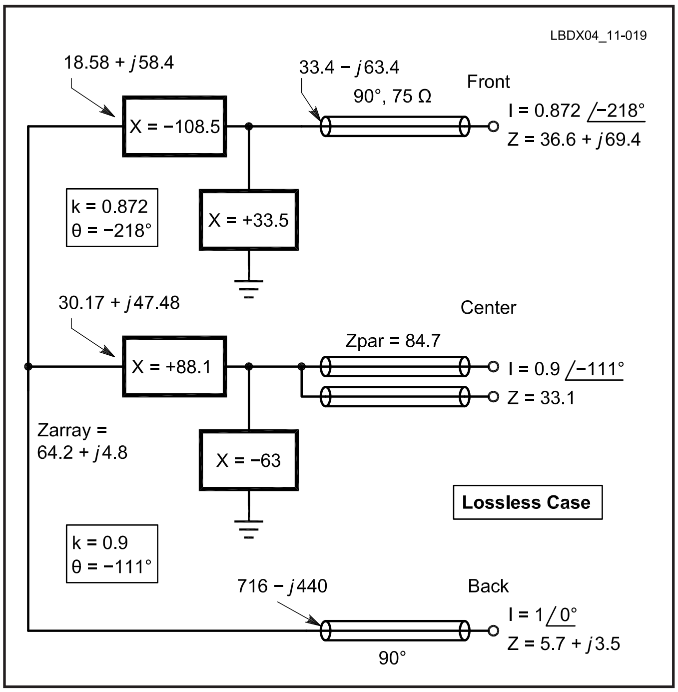



**Fig 11-19—Lahlum/Lewallen feed network for a λ/4-spaced Four-Square array using current-forcing with 75-Ω, λ/4 feed lines and where the back element is directly fed.**

### 3.4.5.12. Using Lahlum's formulas for desired phase angles—the modified Lewallen method

So far we have used θ = 90° in the generic formulas shown in Section 3.4.5. Robye Lahlum, W1MK, developed the formulas that allow us to use the L-network to obtain a phase shift other than 90° with different current magnitudes, and he decided to share them with me for publication in this book, for which I am very grateful!

As we will see in Section 5 it appears that we can significantly improve the performance of a Four-Square by not feeding the element in quadrature (in 90° steps) and with equal current magnitudes. Jim Breakall, WA3FET developed such an optimized version of a Four-Square array.

In **Fig 11-19** the back element is the reference element, with θ = 0° and k = 1. The two center elements are fed with a phase angle of −111° and a current magnitude ratio of k = 0.9, the front element with θ = −218° and k = 0.872. In this example I used the following feed impedances for a full-size quarter-wave spaced Four-Square, including 2 Ω ground-loss resistance:

Z-front element: 36.6 + *j* 69.4 Ω
Z-center-elements: 33.1 Ω
Z-back element: 5.7 + *j* 3.5 Ω

The component values are computed in the *Lahlum.xls* spreadsheet. See **Table 11-4** and **Table 11-5**, based on a 75-Ω cable impedance. If I had used 50-Ω feed lines, the array impedance would have been approx 2.25 times lower than shown in Fig 11-19 or approx 28 – *j* 2.2 Ω.

As explained above, the formulas used in the spreadsheet assume zero-loss transmission lines. In most cases this will give a result accurate enough to tell you what the approximate value of the components of the L-network will be. You can however also do the exercise including cable losses. See Section 3.4.9 and **Fig 11-20**, which shows but one of the many alternative solutions that can be calculated using the *Lahlum.xls* calculation tool.

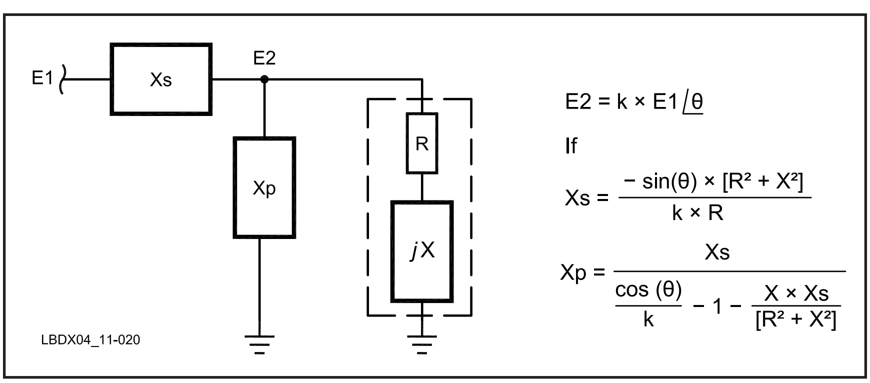



**Fig 11-20—Equations for calculating the L-network components needed to produce a desired phase shift θ, based upon the feed-point impedances (R = real part, X = reactive part). (These equations do not use lossless current-forcing feed lines that are odd multiples of λ/4, although that option is available in the upper portion of the *Lahlum.xls* spreadsheet.) See text for details.**

### 3.4.5.13. Array impedance:

From the spreadsheet we find the input impedance to both L-networks:

Z (to center elements): 30.17 + *j* 47.48 Ω
Z (to front element): 18.58 – *j* 58.4 Ω

Let's now use the COAX TRANSFORMER/SMITH CHART module of the NEW LOW BAND SOFTWARE to calculate the impedances at the end of the λ/4 feed line going to the back element (fed without phasing):

Back element: Z= 5.7 + *j* 3.5 Ω. At the other end of the current-forcing feed

**Table 11-4**
**Lahlum.xls spreadsheet results (see Fig 11-19):**

| | |
|---|---|
| *INPUT DATA* | |
| n elem | 2 |
| Zo | 75.00 Ω |
| R | 33.10 Ω |
| X | 0.00 Ω |
| k | 0.90 |
| theta | –111.00° |
| freq | 3.80 MHz |
| *RESULTS* | |
| Xs | 88.14 Ω |
| Xp | –63.04 Ω |
| Series elem | 3.7 µH |
| Par elem | 664.7 pF |
| Rpar | 104.90 Ω |
| Xpar | 66.65 Ω |
| Rser | 30.17 Ω |
| Xser | 47.48 Ω |

**Table 11-5**
**Lahlum.xls spreadsheet results (see Fig 11-19):**

| | |
|---|---|
| *INPUT DATA* | |
| n elem | 1 |
| Zo | 75.00 Ω |
| R | 36.60 Ω |
| X | 69.40 Ω |
| k | 0.87 |
| theta | –218.00 deg |
| freq | 3.80 MHz |
| *RESULTS* | |
| Xs | –108.51 Ω |
| Xp | 33.46 Ω |
| Series elem | 386.2 pF |
| Par elem | 1.4 µH |
| Rpar | 202.12 Ω |
| Xpar | –64.31 Ω |
| Rser | 18.58 Ω |
| Xser | –58.40 Ω |

line: Z = 716 – *j* 440 Ω. All three in parallel (calculated with the PARALLEL IMPEDANCES module of the NEW LOW BAND SOFTWARE: $Z_{tot}$ = 64.2 – *j* 4.8 Ω.

### 3.4.6. Collins (W1FC) hybrid-coupler method

Fred Collins, W1FC, developed a feed system similar to the Lewallen system in that it uses current-forcing λ/4 feed lines to the individual elements. There is one difference, however. Instead of using an L network, Collins uses a quadrature *hybrid coupler*, shown in **Fig 11-21**.

The hybrid coupler divides the input power (at port 1) equally between ports 2 and 4, with theoretically no power output at port 3 if all four port impedances are the same. When the output impedances are not the same, power will be dissipated in the load resistor connected to port 4. In addition, the phase difference between the signal at ports 2 and 4 will not be different by 90° if the load impedance of these ports is not real or, if complex, they do not have an identical reactive part.

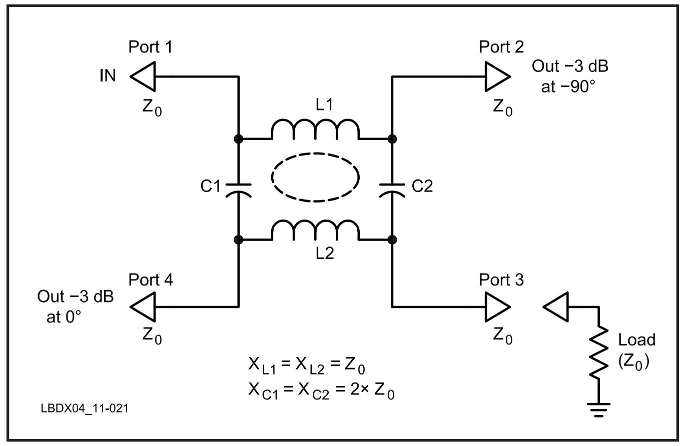



**Fig 11-21—Hybrid coupler providing two –3 dB outputs with a phase difference of 90°. L1 and L2 are closely coupled. See text for construction details.**

We will examine whether or not this characteristic of the hybrid coupler is important to its application as a feed system for a quadrature-fed array.

#### *3.4.6.1. Hybrid coupler construction*

The values of the hybrid coupler components are:

XL1 = XL2 = 50-Ω system impedance
XC1 = XC2 = 2 × 50 Ω= 100 Ω

For 3.65 MHz the component values are:

$$L1 = L2 = \frac{X_L}{2\pi f} = \frac{50}{2\pi \times 3.65} = 2.18\,\mu H$$

$$C1 = C2 = \frac{10^6}{2\pi f X_C} = \frac{10^6}{2\pi \times 3.65 \times 100} = 436\text{ pF}$$

When constructing the coupler, you should take into account the capacitance between the wires of the inductors, L1 and L2, which can be as high as 10% of the required total value for C1 and C2. The correct procedure is to first wind the tightly coupled coils L1 and L2, then measure the inter-winding capacitance and deduct that value from the theoretical value of C1 and C2 to determine the required capacitor value. For best coupling, the coils should be wound on powdered-iron toroidal cores. The T225-2 (µ = 10) cores from Amidon are a good choice for power levels well in excess of 2 kW. The larger the core, the higher the power-handling capability. Consult Table 6-3 in Chapter 6 for core data. The T225-2 core has an $A_L$ factor of 120. The required number of turns is calculated as:

$$N = 100\sqrt{\frac{2.18}{120}} = 13.4\text{ turns}$$

The coils can be wound with AWG #14 or AWG #16 multi-strand Teflon-covered wire. The two coils can be wound with the turns of both coils wound adjacent to one another, or the two wires of the two coils can be twisted together at a rate of 5 to 7 turns per inch before winding them (equally spaced) onto the core.

At this point, measure the inductance of the coils (with an impedance bridge or an LC meter) and trim them as closely as possible to the required value of 2.09 µH for each coil. Do *not* merely go by the calculated number of turns, since the permeability of these cores can vary quite significantly from production lot or one manufacturer to another. Moving the windings on the core can help you fine-tune the inductance of the coil. Now the interwinding capacitance can be measured. This is the value that must be subtracted from the capacitor value calculated above (436 pF). A final check of the hybrid coupler can be made with a vector voltmeter or a dual-trace oscilloscope. By terminating ports 2, 3 and 4 with 50-Ω resistors, you can now fine-tune the hybrid for an exact 90° phase shift between ports 2 and 4. The output voltage amplitudes should be equal.

### 3.4.7. Gehrke (K2BT) method

Gehrke, K2BT, has developed a technique that is fairly standard in the broadcast world. The elements of the array are fed with randomly selected lengths of feed line, and the required feed currents at each element are obtained by the insertion of discrete component (lumped-constant) networks in the feed system. He makes use of L networks and constant-impedance T or pi phasing networks. The detailed description of this procedure is given in Ref 924.

The Gehrke method consists of selecting equal lengths (not necessarily 90°lengths) for the feed lines running from the elements to a common point where the array switching and matching are done. With this method, the length of the feed lines can be chosen by the designer to suit any physical requirements of the particular installation. The cables should be long enough to reach a common point, such as the middle of the triangle in the case of a triangle-shaped array.

As this method is rarely used in amateur circles, I decided not to describe it in detail in this edition of the book (but it was covered in all previous editions). This method however has the tremendous merit that it was the first one described in amateur literature that was technically 100% correct.

### 3.4.8. Lahlum (W1MK)/Gehrke(K2BT)

The Lahlum/Lewallen method described in sect 3.4.6 can be applied with any coax feed length—the length does not necessarily have to be λ/4 or odd multiples of λ/4. While the use of current-forcing is a very desirable feature there are situations where you might not care to use current forcing. For example, the use of the array on multiple bands with the use of the same coax feed for both bands. The Lahlum/Lewallen method is suitable in this situation.

I called this system the "Lahlum/Gehrke" system, since it uses the mathematics developed by Robye Lahlum, W1MK, and follows more or less the principle of Gehrke's original methods, where arbitrary lengths of feed lines were used to the elements.

In this case we will first have to calculate the impedances at the end of the feed lines; eg, using the COAX TRANSFORMER/SMITH CHART module of the NEW LOW BAND SOFTWARE. The formulas involved are given in Fig 11-20. R and *j* X are the impedance values of the feed impedances of the antenna elements, transformed by the coaxial feed line.

In the situation explained in Section 3.4.6, k is the ratio of the feed currents when we use current-forcing feed lines. In this application however, k = E1/E2, the ratio between the *voltages* at the end of the feed lines. These feed lines are not necessarily 90° long—or odd multiple thereof— and they do not even have to be of equal length. θ is again the phase shift caused by the L-network. It is the phase angle difference between the voltages at the end of the two equal-length feed lines. More precisely it is the difference between the voltage phase angle at the output of the L-network and the phase angle at the input of the network. θ *must* be negative. If necessary subtract 360° to obtain a negative value. **Fig 11-22** shows the principle.

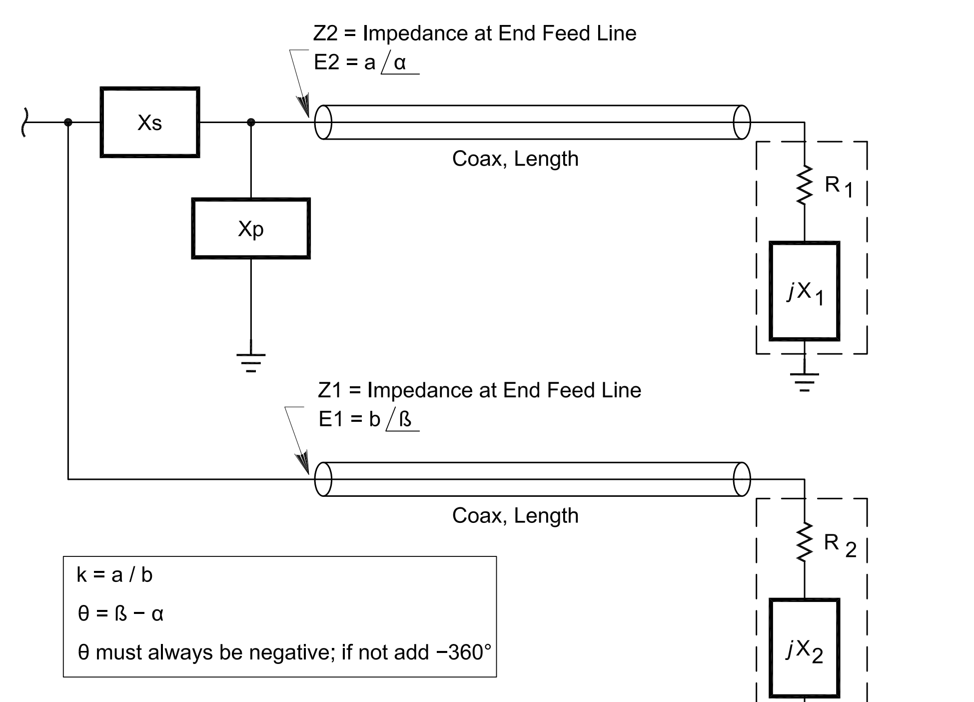



**Fig 11-22—Basic setup for a 2-element array. Starting from the feed impedances of the elements we calculate Z and E at the end of the two feed lines. The schematic shows the requirements for the network. It must transform the voltage a/∝° to b/ß° with a phase shift of θ = (ß°– ∝°), while k = a/b. See text for details.**

This non-current-forcing feed systems is an elegant solution where you want to built two-band arrays; for example, covering 80 meters with wide spacing (approx λ/4) and 160 meters with close (λ/8) spacing. Let's work out an example for a 2 element, λ/8 spacing case, where the phase shift is –135°. Through antenna modeling we obtain the following element impedance values:

Back element:

$I_{back}$= 1 /0° A

$Z_{back}$= 13 - *j* 21 Ω

Front element:

$I_{front}$= 1/-135° A

$Z_{front}$ = 18 + *j* 23 Ω

Using the COAX TRANSFORMER/SMITH CHART module of the NEW LOW BAND SOFTWARE, the values at the end of a 38.4° long feed line are calculated. (Note: It's *not* necessary that both feed lines be of equal length, unless of course you want to switch directions.). I used a frequency of 1.83 MHz, using real cable (RG-213, 0.2 dB loss/100 feet). We now need to look at the voltage at the end of the feed lines, since we need to connect them in parallel (equal voltages required!). The transformed values are:

At end feed line to back element:

Eb' = 18.12 /54.04° V

Zb' = 12.07 + *j* 12.13 Ω

At end feed line to front element:

Ef' = 51.23 /-61.24° V

Zf' = 61.07 – *j* 69.94 Ω

We need to insert an L network in either the feed line to the front or to the back element. This L-network has to perform the followings two tasks:

- Perform the required phase shift
- Perform the required voltage transformation so that the input voltage to the L-network is identical to the voltage at the end of the other feed line (so that we can connect them in parallel).

### *3.4.8.1. Solution 1*

See **Fig 11-23**. We put the L-network in the feed-line going to the

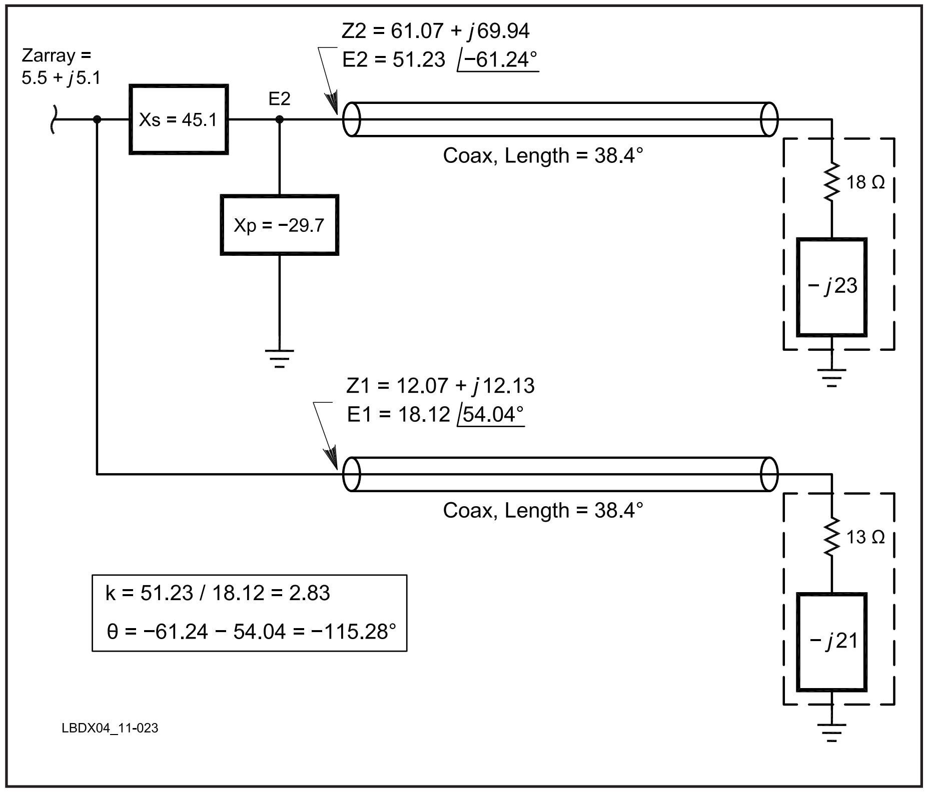



**Fig 11-23—First solution for a 2-element end-fire array (λ/8 spacing). See text for details.**

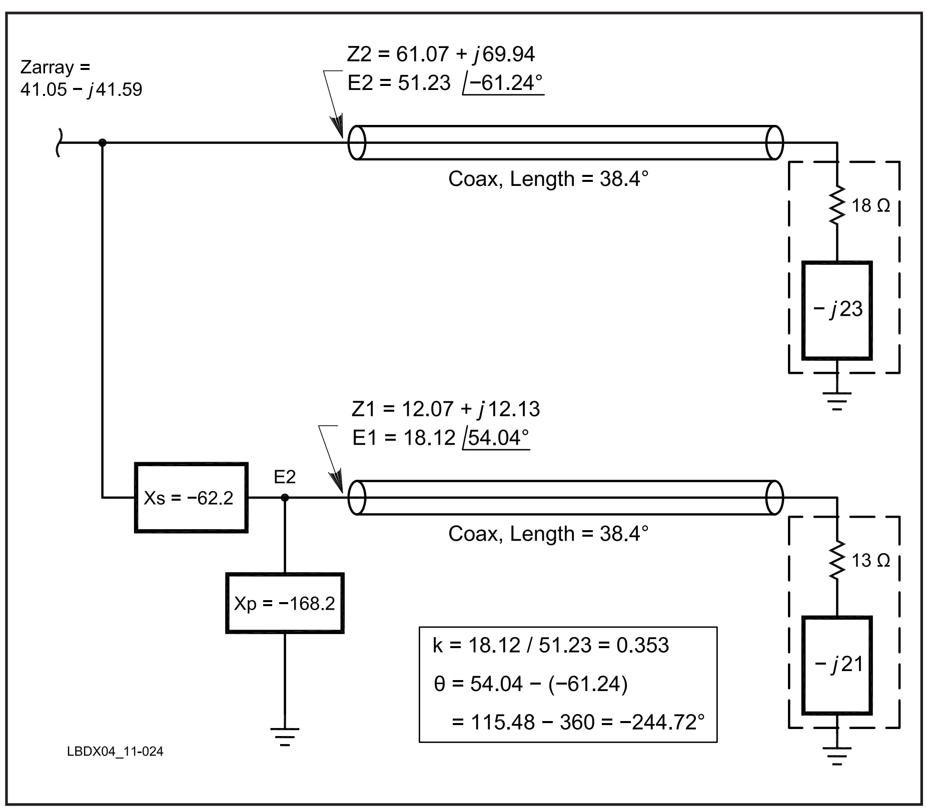



**Fig 11-24—Second, more practical, solution for a 2-element end-fire array (λ/8 spacing). See text for details.**

**Table 11-6**
**Lahlum.xls spreadsheet results (see Fig 11-24):**

| | |
|---|---|
| *INPUT DATA* | |
| R | 12.07 Ω |
| X | 12.13 Ω |
| k | 0.353 |
| theta | –244.80° |
| freq | 3.80 MHz |
| *RESULTS* | |
| Xs | –62.2 Ω |
| Xp | –168.2 Ω |
| Series elem | 673.9 pF |
| Par elem | 249.2 pF |
| Rpar | 194.69 Ω |
| Xpar | –54.06 Ω |
| Rser | 13.94 Ω |
| Xser | –50.19 Ω |

front element. θ is the difference between the voltage phase angle at the output of the L-network and the phase angle at the input of the network. λ must be negative. If necessary subtract 360° to obtain a negative value.

$\theta = (-61.24) - (54.04) = -115.28°$

k = ratio of the voltage magnitudes at the end of the feed lines: k = 51.23/18.12 = 2.83

We can plug these values in the formulas shown in Fig 11-20, or better yet use the special spreadsheet tool *Lahlum.xls*. This tool allows you to calculate the values of the L network directly. For this example (see **Fig 11-23**):

$X_{ser} = 45.11\ \Omega$
$X_{par} = -29.7\ \Omega$

An impedance of –29.7 Ω in parallel with 61.07 – *j* 69.94 Ω gives 10.07 – *j* 36.34 Ω. Adding the series reactance of 45.11 Ω gives 10.07 + *j* 8.76 Ω. Paralleling this impedance with 12.07 + *j* 12.13 Ω gives 5.5 + *j* 5.1 Ω for the array's feed impedance.

### 3.4.8.2. Solution 2

See **Fig 11-24** and **Table 11-6**. The L-network is in the feed line going to the back element:

$\theta = (54.04) - (-61.24) = +\ 115.28 = (-360+115.28) = -244.72°$

k = 18.12/51.23 = 0.353

$X_{ser} = -62.2\ \Omega$

$X_{par} = -168.2\ \Omega$

Note that this requires two capacitors, rather than a capacitor and an inductor, for the L-network. –168.2 Ω in parallel with 12.07 + *j* 12.13 Ω gives 13.94 + *j* 12.00 Ω. Adding the series reactance of –62.2 Ω gives 13.93 – *j* 50.2 Ω. Paralleling this impedance with 61.07 + *j* 69.94 Ω gives 47.04 – *j* 41.59 Ω for the array feed impedance

Both solutions are valid, the only difference is the resulting input impedance. In Solution 1 the resulting input impedance is very low (5.5 + *j* 5.1 Ω). Solution 2 yields an array feed impedance that is much closer to 50 Ω (47 – *j* 41 Ω), and the use of a series inductor would give an almost perfect match to 50-Ω cable.

This approach to solving the problem of obtaining the correct amplitude and phase shift using coax feeds of any length is similar to the method of Gehrke, K2BT, however it results in much fewer circuit elements. Solving this same problem using Gehrke's method would result in the need for six or seven elements (see *Low Band DXing*, Editions 1, 2 or 3) , all of which would affect the amplitude/phase relationships.

Using the Lahlum/Lewallen approach, four elements in general would be required. Two of them would be an L-network matching the array input impedance to the feed-line impedance and only two of them affect the amplitude/phase relationship, thus making it much easier to adjust.

### 3.4.8.3. Adjusting the network values

If you do *not* use current-forcing (feed lines that are λ/4 or odd multiples thereof), you cannot use the testing and adjustment procedure as described in Section 3.6.2. (measuring voltages at the end of the feed lines). In this case you will have to use a small current probe at the elements (see Section 3.5.5. and Fig 11-29).

### 3.4.8.4. Other applications of the software

While the calculation procedures described in Section 3.4.5 assume current-forcing feed lines without losses, you can use the above procedure to take actual losses into account. You first need to calculate the impedances at the end of the current forcing feed lines, using the COAX TRANSFORMER/SMITH CHART module of the NEW LOW BAND SOFTWARE (option "with cable losses") and then use these values as input date for the *Lahlum.xls* spreadsheet.

## 3.4.9. Choosing a feed system

Until Gehrke published his excellent series on vertical arrays, it was general practice to simply use feed lines as phasing lines, and to equate electrical line length to phase delay under all circumstances. We now know that there are better ways of accomplishing the same goal (Ref Section 3.3.1).

Fortunately, as Gehrke states, these vertical arrays are relatively easy to get working. **Fig 11-25** shows the results of an analysis of the 2-element cardioid array with deviating feed currents. The feed-current magnitude ratio as well as the phase angle are quite forgiving so far as gain is concerned. As a matter of fact, a greater phase delay (eg, 100° versus 90° will increase the gain by about 0.3 dB. The picture is totally different so far as F/B ratio is concerned. To achieve an F/B of better than 20 dB, the current magnitude as well as the phase angle need to be tightly controlled. But even with a "way off" feed system it looks like you always get between 8 and 12 dB of F/B ratio, which is indeed what we used to see from arrays that were incorrectly fed with coaxial phasing lines having the electrical length of the required phase shift.

### 3.4.9.1. Collins (hybrid coupler) system

We know that the perfect 90° phase shift with identical antenna feed-current magnitudes can never be obtained with this system because the hybrid is never terminated in its design impedance (50 Ω) but rather in different complex feed impedances of the elements of the array.

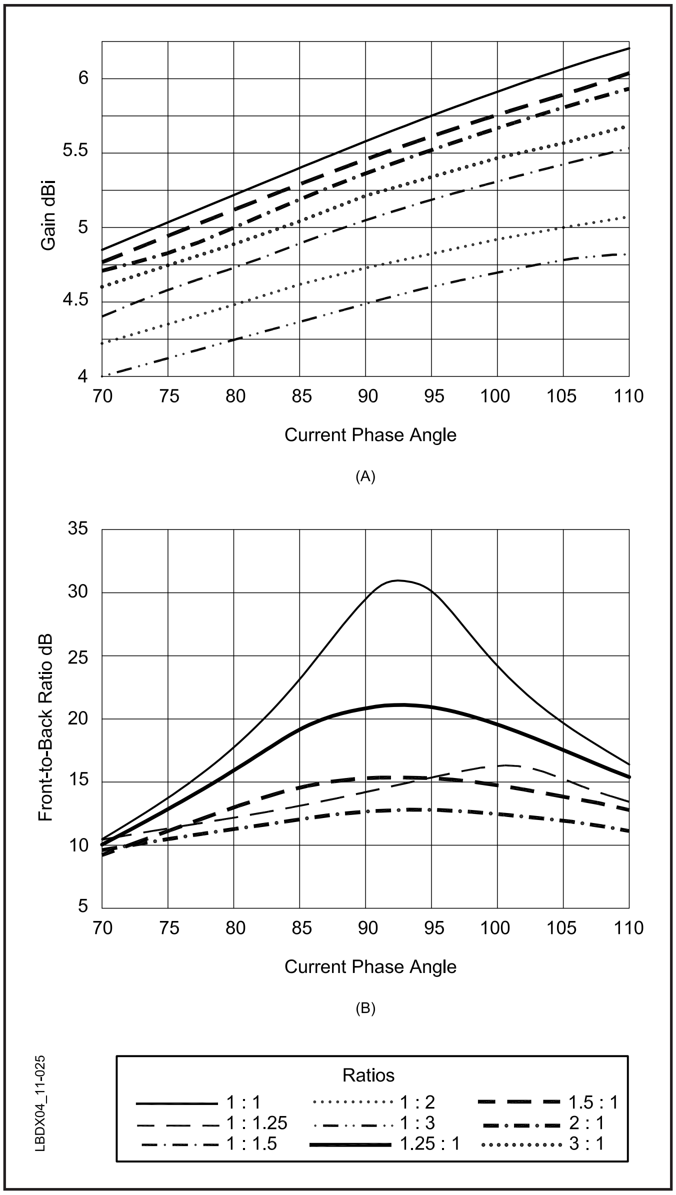



**Fig 11-25—Calculated gain and the front-to-back ratio of a 2-element cardioid array versus current magnitudes and phase shifts. Calculations are for very good ground at the main elevation angle. The array tolerates large variations so far as gain is concerned, but is very sensitive so far as front-to-back ratio is concerned.**

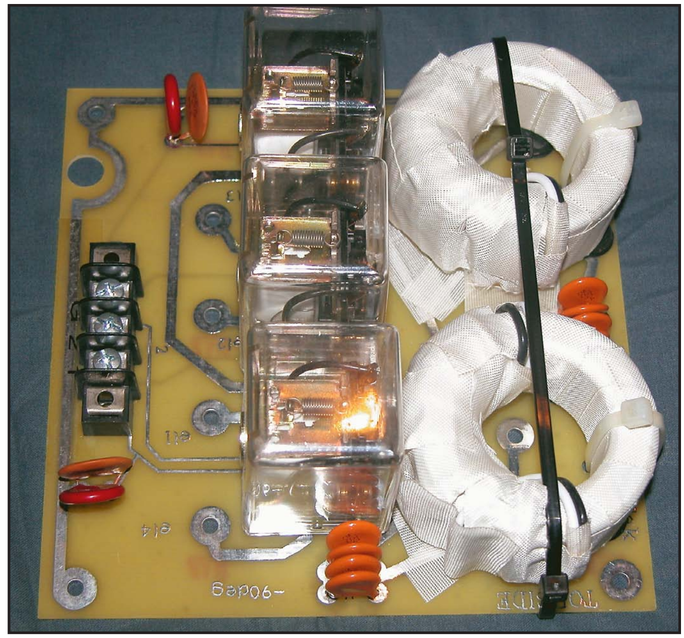


**Fig 11-26—The internal works of the Comtek hybrid coupler: PC board showing two large toroidal cores: one is used in the hybrid coupler, while the second one serves to make a 180° phase-reversal transformer (used instead of a 180° phasing line). Note also the three heavy-duty relays for direction switching.**

### *3.4.9.1.1. Performance of the hybrid coupler*

I have tested the performance of a commercially made hybrid coupler (Comtek, see Section 3.4.6 and **Fig 11-26**). First the coupler was tested with the two load ports (ports 2 and 4) terminated in a 50-Ω load resistor. Under those conditions the power dissipated in the 50-Ω dummy resistor (port 3) was 21 dB down from the input power level. This means that the coupler has a directivity of 21 dB under ideal loading conditions (equivalent to 12 W dissipated in the dummy load for a 1500-W input). The results were identical for both 3.5 and 3.8 MHz. The input SWR under the same test conditions was approximately 1.1:1 (a 25-dB return loss).

I also checked the hybrid coupler for its ability to provide a 3-dB signal split with a 90° phase-angle difference. When the two hybrid ports were terminated in a 50-Ω load I measured a difference in voltage magnitude between the two output ports of 1.7 dB, with a phase-angle difference of 88° at 3.8 MHz. At 3.5 MHz the phase-angle difference remained 88°, but the difference in magnitude was down to 1.2 dB. Theoretically the difference should be 0 dB and 90°. A 1.2-dB difference means a voltage or current ratio of $k = 0.87$.

The commercially available hybrid coupler system from Comtek Systems (**comtek4@juno.com**) uses a toroidal-wound transmission line to achieve a 180° phase shift over a wide band-width. The phase transformer consists of a bifilar-wound conductor pair, where wire A is grounded on one end and wire B on the other end of the coil. The other two ends are the input and output connections, whereby the voltages are shifted 180° in phase. This approach eliminates the long (λ/2) coax that is otherwise required for achieving the 180° phase shift and it is broadbanded as well.

I measured the performance of this "compressed" 180° phasing line. Using a 50-Ω load, the output phase angle was –168°, with an insertion loss of 0.8 dB. With a complex-impedance load the phase shift varied between –160° and –178°. Measurements were done with a Hewlett-Packard vector voltmeter. The hybrid coupler was also evaluated using real loads in a Four-Square array.

After investigating the components of the Comtek hybrid-coupler system, I evaluated the performance of the coupler (without the 180° phase inverter transformer), using impedances found at the input ends of the λ/4 feed lines in real arrays as load impedances for ports 2 and 4 of the coupler. Let us examine the facts and figures for our 2-element end-fire cardioid array.

The SWR on the quarter-wave feed lines to the two elements (in the cardioid-pattern configuration) is not 1:1. Therefore, the impedance at the ends of the quarter-wave feed lines will depend on the element impedances and the characteristic impedances of the feed lines. We want to choose the feed-line impedances such that a minimum amount of power is dissipated in the port-3 terminating resistor.

The impedances at the end of the 90°-long real-world feed lines (λ/4 RG-213 with 0.35 dB/100 feet on 80 meters) are:

$Z1' = 42.81 - j\,16.18\ \Omega$
$Z2' = 63.1 - j\,56.94\ \Omega$

These values are reasonably close to the 50-Ω design impedance of our commercial hybrid coupler. With 75-Ω feed lines the impedance would be:

$Z1' = 95.11 - j\,35.88\ \Omega$
$Z2' = 141.05 - j\,125.4\ \Omega$

It is obvious that for a 2-element cardioid array, 50 Ω is the logical choice for the feed-line impedance. This can be different for other types of arrays. The basic 4-element Four-Square array, with λ/4 spacing and quadrature-fed, is covered in detail in Section 4.7. A special version of the Four-Square array is analyzed in detail in Section 6.

### *3.4.9.1.2. Array performance*

Although the voltage magnitudes and phase at the ends of the two quarter-wave feed lines are not exactly what is needed for a perfect quadrature feed, it turns out that the array only suffers slightly from the minor difference. The incorrect phase angle will likely deteriorate the F/B, but the gain will remain almost the same as with the nominal quadrature driving conditions, which again, do not result in optimum gain nor directivity (see also Fig 11-25).

### *3.4.9.1.3. Different design impedance*

We can also design the hybrid coupler with an impedance that is different from the 50-Ω quarter-wave feed-line impedance in order to realize a lower SWR at ports 2 and 4 of the coupler. The load resistor at port 3 must of course have the same ohmic value as the hybrid design impedance. Alternatively we can use a standard 50-Ω dummy load with a small L network connected between the load and the output of the hybrid coupler.

With the aid of the software module SWR ITERATION, you can scan the SWR values at ports 2 and 4 for a range of design impedances. The results can be cross-checked by measuring the power in the terminating resistor and alternately connecting 50-Ω and 75-Ω quarter-wave feed lines to the elements. A practical design case is illustrated in Section 4.7.1.2.

By choosing the most appropriate feed-line impedance as well as the optimum hybrid-coupler design impedance, it is possible to reduce the power dissipated in the load resistor to 2% to 5% of the input power. Whether or not reducing the lost power to such a low degree is worth all the effort may be questionable, but covering the issue in detail will certainly help in better understanding the hybrid coupler and its operation as a feed system for a phased array with elements fed in quadrature.

### *3.4.9.1.4. Bottom line*

The Collins feed method (with the hybrid coupler) is only applicable in situations where the elements are fed in quadrature relationship (in increments of 90°). We also must realize that the hybrid-coupler system does *not* produce the exact phase-quadrature phase shift unless some very specific load conditions exist (resistive loading or loading with identical reactive components on both ports).

Fortunately most of the quadrature-fed arrays are quite lenient, tolerating a certain degree of deviation from the perfect quadrature condition. We know however that the quadrature feed configuration is not the best configuration, and 0.6 up to 1 dB more gain and better directivity (narrower forward lobe) can be obtained with other phasing angles and different current magnitudes (see Section 4.7.2 and Section 6.8.4).

Over the years the Collins method has become the most popular feed method, clearly because it is a "plug and play" type solution, which works most of the time! The tradeoff for this is that you are not getting peak performance, such as can be obtained with a properly adjusted Lahlum/Lewallen feed system.

One advantage with the Collins system, however, is that essentially the same (but compromised) front-to-back ratio can be achieved over the entire band (3.5 to 4.0 MHz on 80 meters).

Watch out though and don't make the error to judge the operational bandwidth of the hybrid-coupler system by measuring the SWR curve at the input of the coupler. The coupler will show a very flat SWR curve (typically less than 1.3:1) under *all* circumstances, even from 3.5 to 4 MHz or from 1.8 MHz to 1.9 MHz on 160 meters. The reason is that, away from its design frequency, the impedances on the hybrid ports will be extremely reactive, resulting in the fact that nearly all power fed into the system will be dissipated in the dummy resistor. It is typical that an array tuned for element resonance at 3.8 MHz will dissipate 50% to 80% of its input power in the dummy load when operating at 3.5 MHz. The exact amount will depend on the Q factor of the elements. On receive, the same array will still exhibit excellent directivity on 3.5 MHz, but its gain will be down by 3 to 7 dB from the gain at 3.8 MHz, since it is wasting 50% to 80% of the received signal as well into the dummy resistor.

It is clear that the only bandwidth-determining parameter is the power wasted in the load resistor. So stop bragging about your SWR curves, but let's see your dummy-load power instead! The hybrid coupler has the drawback of wasting part of the transmitter power (and receive power as well, but that's probably much less relevant) in the dummy-load resistor. Ten percent power loss may not seem a lot, but on 160 meters, where signals are often riding on or in the noise, 10% of power, which equals 0.5 dB, can be meaningful.

### ***3.4.9.2 Christman system***

The Christman method makes maximum use of the transformation characteristics of coaxial feed lines, thus minimizing the number of discrete components required in the feed network. This is an attractive solution, and should not scare off potential array builders. For a 2-element cardioid array this is certainly a good way to go. Of course, you need to go through the trouble of measuring the impedances.

With arrays of more elements, it is likely that identical voltages will only be found on two lines. For the third line, lumped-constant networks will have to be added. In such case the Lewallen or Lahlum/Lewallen method is preferred.

### *3.4.9.3. Lewallen and Lahlum/Lewallen systems*

#### *3.4.9.3.1. The quadrature Lewallen system*

The Lewallen feed system has been used very successfully by many array builders, especially those that want no compromises and only care for peak performance. The system can produce the right phase angle and feed current magnitude for any load impedance, and one can adjust ("tune") the values of the L-network to obtain the desired values.

Lewallen, W7EL, published in the last several issues of *The ARRL Antenna Book* a number of L-network values for the 2-element cardioid and the 4-element square arrays, which a builder can use for building the L network without doing any measuring.

#### *3.4.9.3.2. Any phase angle with Lahlum's approach*

With Lahlum's introduction in this book of the extra feature that allows you to program any phase angle at any feed current magnitude, the enhanced Lahlum/Lewallen system should be considered as the best engineering choice, and should attract all those who want nothing but the best. Lahlum made the mathematics and the calculation method for this fully flexible system available to all home-builders.

It is interesting however to see that the only commercially available feed-system according to the Lewallen feed system (**www.arraysolutions.com**) in fact already was using an approach that seems to be similar if not identical to the Lahlum system. See **Fig 11-27**.

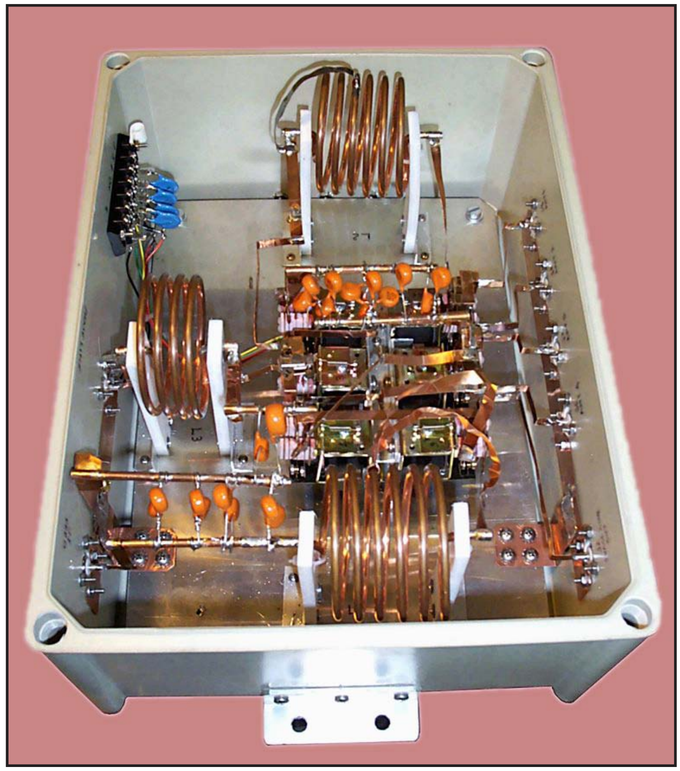


**Fig 11-27—Lahlum/Lewallen feed system for a Four-Square built by Array Solutions (WXØB). The unit includes an L-network for a perfect match to the feed line as well as an omnidirectional position.**

Array Solutions advertises two versions of their Four-Square feed system. One is the quadrature system (0°,–90°, –180°), the other one is called the "optimized version" with phase angles of 0°, –111° and –218°, with unspecified feed-current magnitudes. In the optimized version, the phase in the front element could be made longer than 180° (obtained through a λ/2 feed line) with the addition of a small L-network, which is exactly what is done in Lahlum's solution. Array Solutions tunes all of its feed systems for the desired feed current (magnitude and phase angle).

This system, which employs two L-networks, is "fully adjustable," which is a great advantage. Using quarter-wave (or 3λ/4 feed lines) to your array elements, you can measure the voltage (magnitude and phase) at the start of these lines, and tune the L-networks elements until you obtain exactly what you want. A simple procedure to do that is outlined in Section 3.6.

#### *3.4.9.3.3. My experience*

After having used the hybrid system for a number of years I installed a feed system according the Lahlum/Lewallen system, manufactured by Array Solutions, as shown in Fig 11-27. In Section 3.5 I cover some test equipment I used for tuning the array. See also Chapter 7. When properly tuned, using the right test equipment, you can expect a little better performance from this system compared to the hybrid-coupler system.

The design parameters for my particular Four-Square (using one elevated radial, as described in Section 6) were:

Front element: I = 1.5/–220° A
Center elements: I = 1/–111° A
Back element: I = 0.85/0° A

These feed currents give about 0.6 dB more gain than a perfectly working quadrature feeding solution and the directivity is much enhanced (see also Section 3). I used a vector voltmeter to measured the voltages at the start of the λ/4 feed lines, and transformed to the feed currents at the elements. The measurement was confirmed used the method described in Section 3.6. A multi-channel scope brought further confirmation. The design phase and amplitude are obtained through carefully adjusting the network.

#### *3.4.9.3.4. Bottom line*

I went into great detail in the foregoing sections to explain step by step how you can calculate the Lahlum/Lewallen feed system and build one yourself. The procedure is simple:

- Model the planned array as accurately as possible.
- Use the spreadsheet program (*Lahlum.xls*) to calculate the L-network components.
- Use the NEW LOW BAND SOFTWARE to calculate the array feed impedance.

This is all pretty straightforward. Once you understand, you can calculate any array in less than 10 minutes! Make sure you calculate the L network component values based on real antenna impedances and not 50 Ω. This would yield incorrect values.

## 3.5. Measuring and Tuning

### 3.5.1. Can I put up an array without any test equipment?

None of the arrays described in this chapter can be built or set-up without any measuring. The simplest array uses a quadrature configuration, which makes it possible to use a hybrid coupler for obtaining the required phase shift within most often acceptable tolerances. Even in that case, the elements of the array will have to be tuned to proper resonance. Use an SWR meter to trim the elements to resonance. Don't forget to decouple the "other" elements. Just assume the point of lowest SWR is the resonant frequency (which is not quite true) , and you will be close enough for a 2, 3 or 4 element array fed (in quadrature) with a hybrid coupler. The only other thing you should measure in such an array is the power dumped in your hybrid termination resistor. This should never be more than about 10% of the power going into the hybrid. If the power is high, try 75-Ω, λ/4 feed lines instead of 50-Ω lines, or vice versa. OK, so far we have not needed any special test equipment!

In order to obtain maximum directivity from an array, it is essential that the self-impedances of the elements be identical. Measurement of these impedances requires special test equipment, and the method explained in Section 3.6 is recommended. Equalizing the resonant frequency can be done by changing the radiator lengths, while equalizing the self-impedance can be done by changing the number of radials used. If you start putting down perfectly identical and symmetrical radial systems, you will likely get very similar values for the resistive part of the various elements. If you cannot easily get equal impedances, you will have to suspect that one or more of the array elements are coupling into another antenna or conducting structure. Take down all other antennas that are within λ/2 from the array to be erected. Do not change the length of one of the radiators to get the equal values for the resistive parts of the elements. The elements should all have the same physical height (within a few percent).

### 3.5.2. Can I cut my λ/4 feed lines without special test equipment?

Yes you can, but first a word of warning: Never go by the published velocity-factor figures, certainly not when you are dealing with foam coax. There are several valid methods for cutting λ/4 or λ/2 cable lengths.

You can simply use your transceiver, a good SWR bridge and a good dummy load to cut your phasing lines. Maybe the accuracy will not be as good as with other methods described later, but it is totally feasible. Connect your transmitter through a good SWR meter (a Bird 43 is a good choice) to a 50-Ω dummy load. Insert a coaxial-T connector at the output of the SWR bridge. See **Fig 11-28**.

If you need to cut a quarter-wave line (or an odd multiple of λ/4), first short the end of the coax. Make sure it is a good short, not a short with a lot of inductance. Insert the cable in the T connector. If the cable is a quarter-wave long, the cable end at the T connector will show as an infinite impedance and there will be no change at all in SWR (will remain 1:1). If you change the frequency of the transmitter you will see that on both sides of the resonant frequency of the line, the SWR will rise rather sharply. For fine tuning you can use high power (eg, 1 kW) and use a sensitive meter position for measuring the reflected power. I have found this method very accurate, and the cable lengths can be trimmed very precisely.

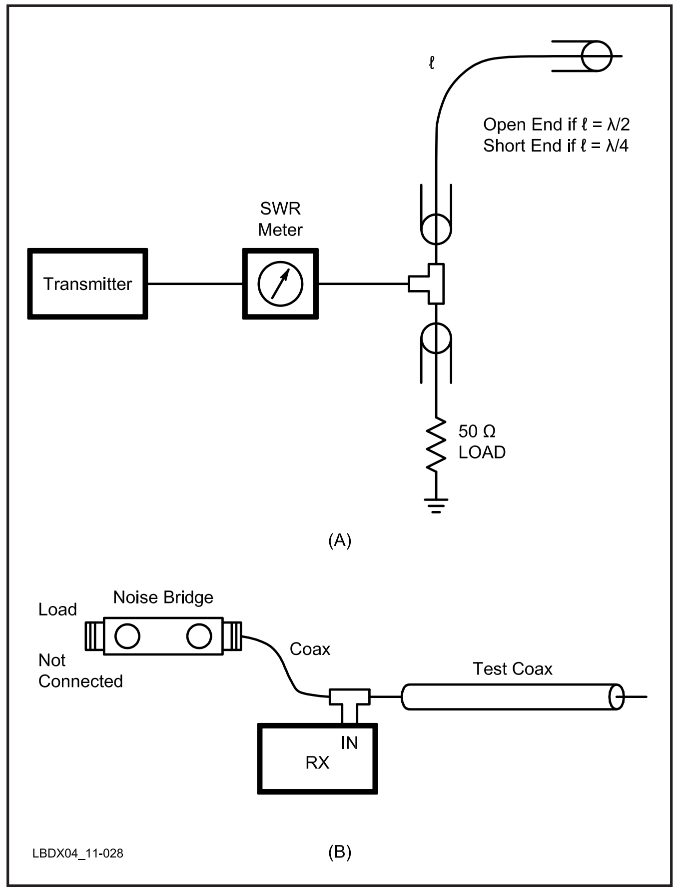



**Fig 11-28—At A, very precise trimming of λ/4 and λ/2 lines can be accomplished by connecting the line under test in parallel with a 50-Ω dummy load. Watch the SWR meter while the line length or the transmitter frequency is being changed. At B, alternative method uses a noise bridge and a receiver. See text for details.**

Make sure the harmonic content from your transmitter is very low. It's a good idea to use a good low-pass or bandpass filter between the generator (transmitter) and the T-connector. A W3NQN bandbass-filter (see Chapter 15, Section 6.3) is ideal for this purpose.

### 3.5.3. Is there a better way to cut the λ/4 lines?

Yes, there are more accurate ways:

- Using a noise bridge (two methods are described in Section 3.6.)
- Using your antenna analyzer
- Preferred method: using the W1MK 6-dB hybrid and detector/power meter (see Section 3.6.5.)

### 3.5.4. What about arrays using the other feed systems (Christman/Lewallen Lahlum)?

In this case we do need to measure the *self impedance* of the elements. This means you need some test equipment.

- You can use your MFJ or AEA antenna analyzer, but their precision is not always very good.
- Much better is to use the W1MK method described in Section 3.6.4.
- Best is to use a professional network analyzer or the VNA (Vector Network Analyzer) described is Section 3.6.9.
- Or use a good old-fashioned Impedance Bridge (eg, General Radio) as described in Section 3.6.10.

You should not only measure the self impedances, but you should try to make them equal, as explained in Section 3.5. Once you have measured the self impedances of all elements, you can calculate the feed impedances, as explained in Section 3.3. Check if the values you calculated are in the same ballpark as the results you obtained through modeling.

If you use a Christman feed method you should now look for points on both feed lines where the voltages are identical (see Section 3.4.2). If you use a Lewallen/Lahlum feed system, you can now calculate the value of the L networks(s) using the *Lahlum.xls* spreadsheet tool, as explained in Sections 3.4.5.

### 3.5.5. How can I measure that the values of the feed-current magnitude and phase angle at the elements are what I really want?

It is essential to be able to measure the feed current to assess the correct operation of the array. A good-quality RF ammeter is used for element-current magnitude measurements and a good dual-trace oscilloscope to measure the phase difference. The two inputs to the oscilloscope will have to be fed via identical lengths of coaxial cable.

**Fig 11-29** shows the schematic diagram of the RF current probes for current amplitude and phase-angle measurement. Details of the devices can be found in Ref 927. D. M. Malozzi, N1DM, pointed out that it is important that the secondary of the toroidal transformer always sees its load resistor, as otherwise the voltage on the secondary can rise to extremely high values and can destroy components and also the input of an oscilloscope if the probe is to be used with a scope. He also pointed out that it is best to connect two identical load resistors at each end of the coax connecting the probe to the oscilloscope. Both resistors should have the impedance of the coax. Make sure the resistors are non-inductive, and of adequate power rating. It is not necessary to do your measurement with high power (nor advisable from a safety point of view).

### 3.5.6. Are there other methods that are more accurate?

Measuring voltage magnitude and phase is easier than measuring current magnitude and phase. We learned in Section 3.4.5 that λ/4 feed lines have this wonderful property called current-forcing. The property allows us to measure voltage at one end of a λ/4 cable to tell us the current at the other end of that cable. This means we make our feed lines quarter-wave (or 3-quarter-wave), and measure the voltage at the end of the feed lines where they all come together.

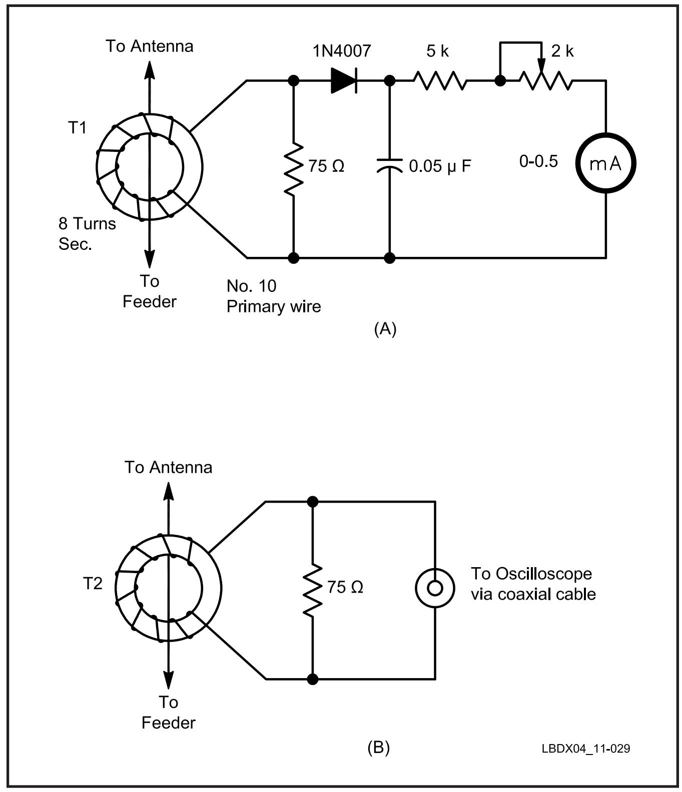



**Fig 11-29—Current amplitude probe (at A) and phase probe (at B) for measuring the exact current at the feed point of each array element. See text for details.**

**T1, T2—Primary, single wire passing through center of core; secondary, 8 turns evenly spaced. Core is ½-in. diameter ferrite, $A_L$ = 125 (Amidon FT-5061 or equivalent).**

### 3.5.7. How do I measure magnitude and phase of these voltages?

The HP Vector-voltmeter (model HP-8405A) is an ideal tool, provided you can find one that has a probe in good condition. Surplus HP-8405As very often have defective probes!

### 3.5.8. Do I really need such lab-grade test-equipment?

No, a very attractive, simple and inexpensive, but very accurate, test method is described in Section 3.6.2.

## 3.6. Test Equipment and Test Procedures for Array Builders

### 3.6.1. Dual channel RF detector/wattmeter (by W1MK)

Various test methods described in this chapter require a sensitive null detector. In most cases a receiver can be used, but a small dedicated and calibrated (in dBm) test instrument is a real asset for any ham who wants to venture into array building.

Robye Lahlum, W1MK, built a dual-channel detector/wattmeter (a modified W7ZOI design), using two AD8307 logamps that give him a sensitivity of better than –70 dBm.

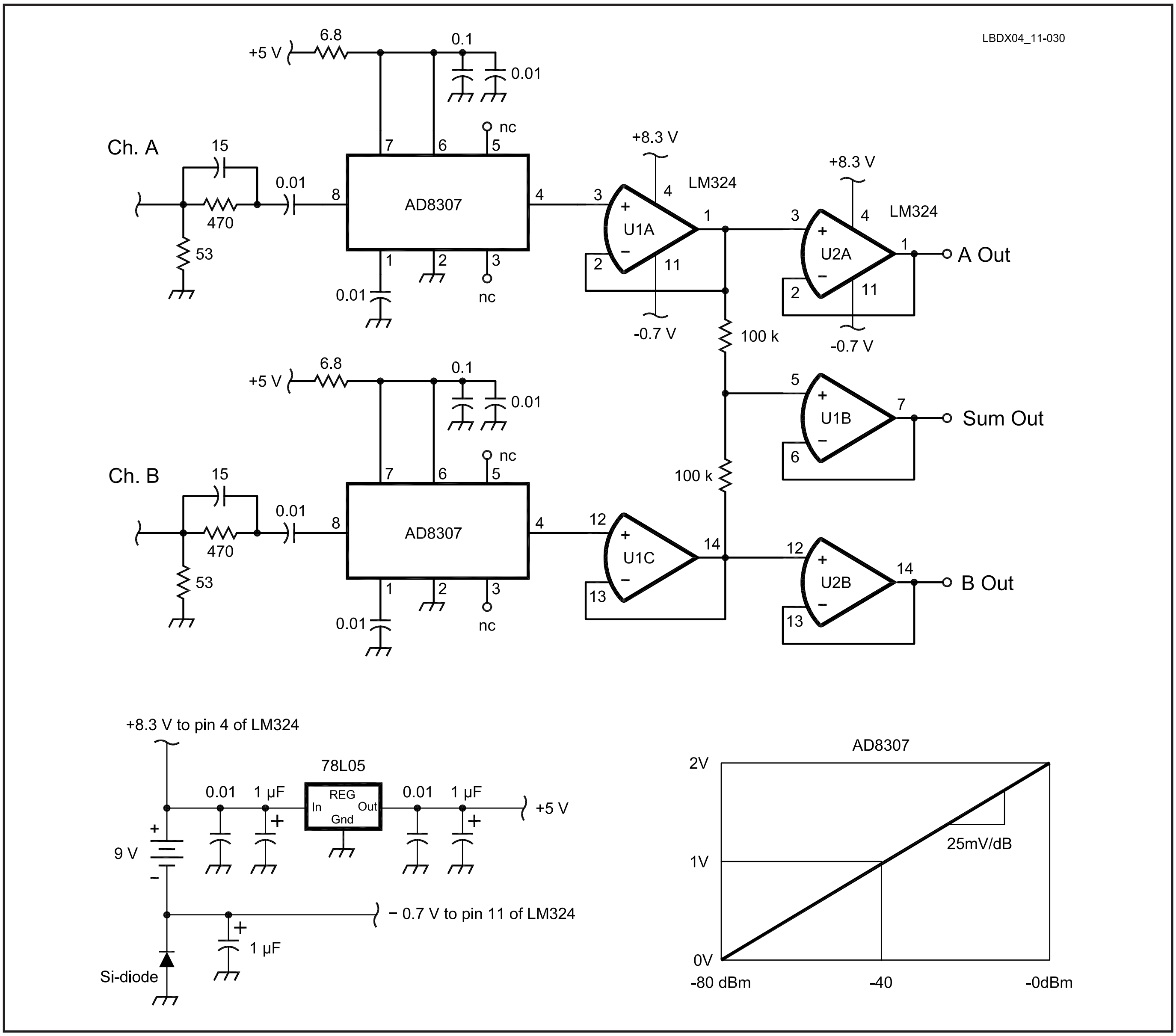



**Fig 11-30—Schematic circuit of the W1MK detector/power-meter circuit. First connect one input and adjust the RF drive for 2 V output. Then the components of the LC circuit(s) are adjusted until the sum output (A + B) reads minimum.**

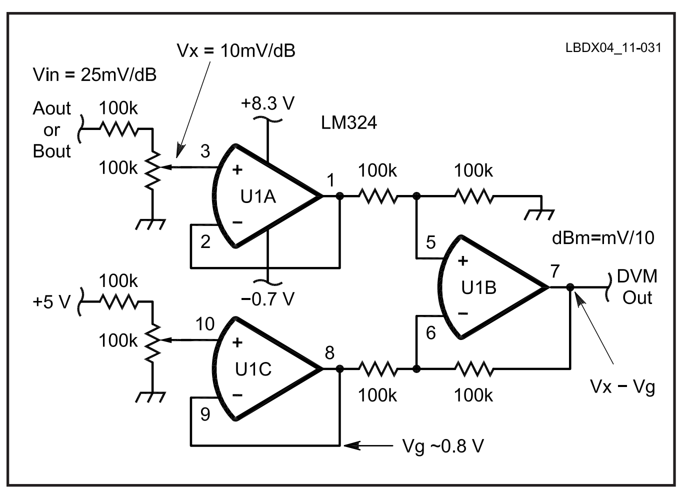



**Fig 11-31—With this additional circuit, the output reading becomes easy to interpret: –50 dBm = –500 mV, and 0 dBm equals 0 mV. If you use a digital voltmeter as an output device, a reading of 0.375 V means a signal of –37.5 dBm.**

The schematic is shown in **Fig 11-30**. In this circuit we see two identical detector/amplifiers, with three outputs: one for channel A, one for channel B and one for the sum of channel A and B. This comes in very handy if we when adjusting a Four-Square array using the Lewallen/Lahlum feed methods using two independent L-networks (see Section 3.6.2).

The output of all three ports varies between 0 and 2 V, where 2 V equals 0 dBm and 0 V equal –80 dBm. The maximum sensitivity is about –75 dBm and it has a bandwidth of approximately 500 MHz.

The circuit shown in **Fig 11-31** makes it possible to read the power in dBm on the scale of the DVM used as indicator.


**Fig 11-32—W1MK's array alignment setup. An Autek Research RF-1 is used as the RF generator. On the left the dual-channel RF wattmeter described in Fig 11-30 and 11-31. The DVM is used as a digital readout.**

The scaling is as follows: Power in dBm = mV/10. Example:

Power in = –50 dBm → –500 mV
Power in = –35 dBm → –350 mV
Power in = 0 dBm → 0 mV

**Fig 11-32** shows W1MK's test setup in action on 80 meters, with an Autek RF-1 used as a signal generator.

### 3.6.2 A hybrid-coupler phase-measuring circuit

The hybrid coupler as used in the W1FC feed systems can be used as the heart of a simple but very effective phase-measuring device for quadrature-fed arrays. If two voltages of identical magnitude but 90° out-of-phase are applied, the bridge circuit will be fully balanced and the output is null. The design also comes from Robye Lahlum, W1MK (Ref 968). **Fig 11-33** shows the hybrid in a simple test circuit for a quadrature-fed Four-Square. After having built the hybrid for the test circuit (see Fig 11-21), use the layout described in **Fig 11-34** to test the hybrid.

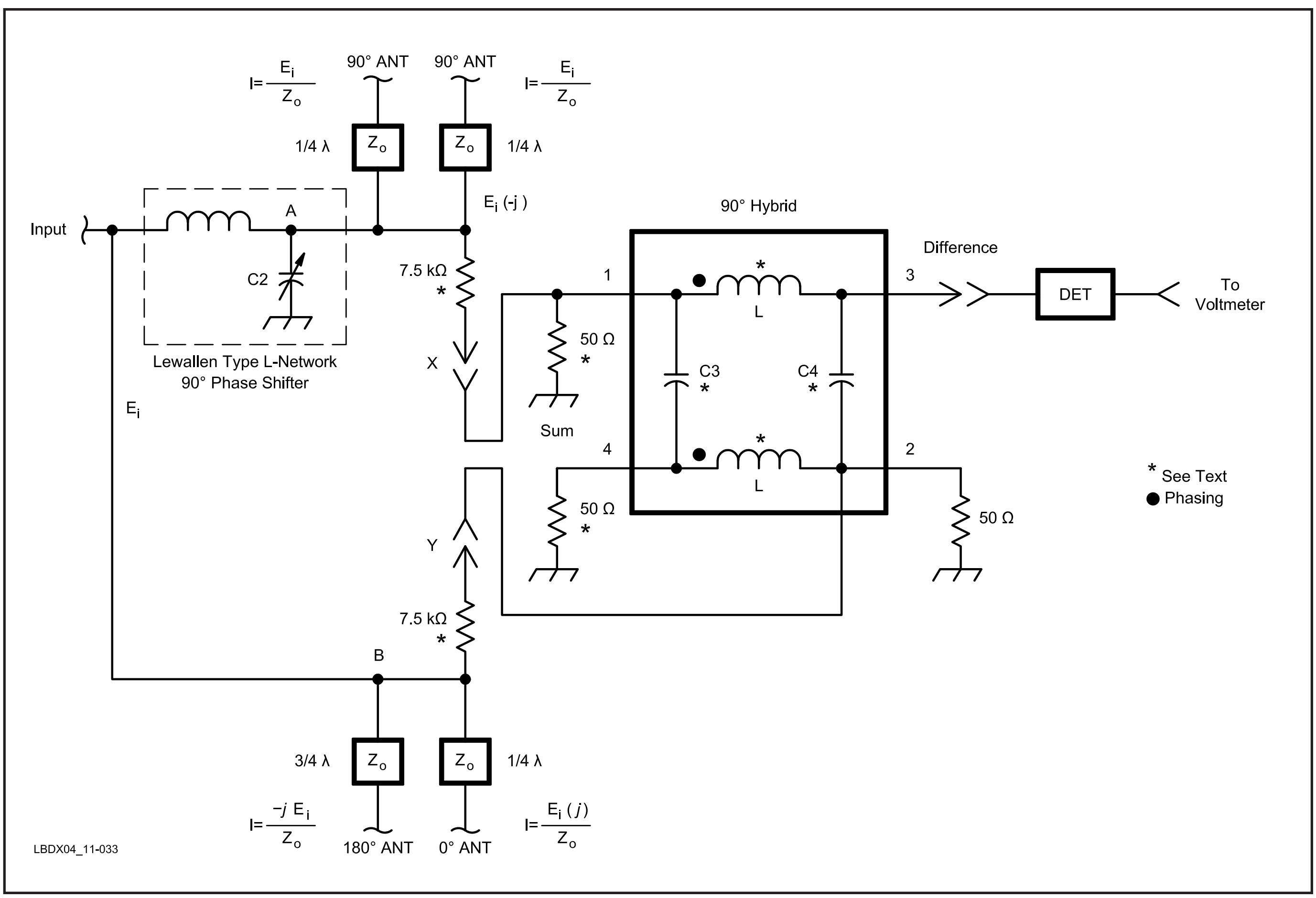



**Fig 11-33—The W1MK phase-measuring setup for quadrature-fed arrays. The unit employs a hybrid coupler as used in the Collins feed system for arrays. The unit can be left permanently in the circuit if the voltage dividing resistors are of adequate wattage. See text for details.**

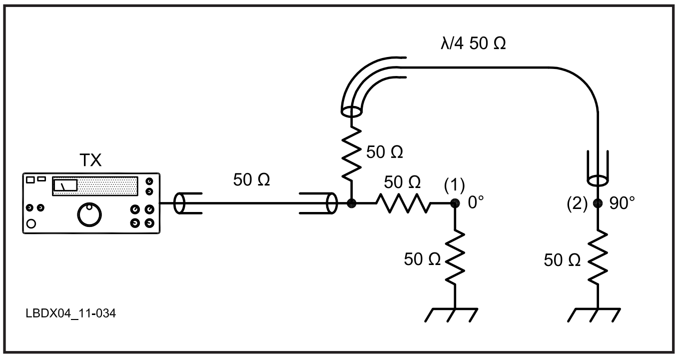



**Fig 11-34—In this phase-calibration system for the quadrature tester, RF voltage from the transmitter is divided down with two 50-Ω series resistors (to ensure a 1:1 SWR), routed directly to a 50-Ω lead, and through a 90°-long 50-Ω line (RG-58) to the second 50-Ω load. For a frequency of 3.65 MHz, the cable has a nominal length of 44.49 feet (13.56 meters). The cable length should be tuned using the method described in Chapter 6 on feed lines and matching.**

Note that the principle can be used with phase angles differences other than 90° as well. Let's work with an example. **Fig 11-35** shows the WA3FET Four-Square, described in Section 4.7. The elements are fed via λ/4 feed lines, which means we can measure the voltages at the end of these lines to determine the currents at the antenna feed point (current equals voltage divided by feed line impedance).

Using a voltage divider (with a high enough dividing ratio so as not to disturb the impedance involved), we sample some voltages at those points and bring them with equal length coaxial cables to our hybrid-coupler test setup. Three possibilities exist:

- Assume first that the array is fed in 90° increments (quadrature feeding). The sampled voltage at the end of our probe lines will be 90° out-of-phase and the output of the hybrid coupler will be zero.
- Assume that we are feeding with 90° phase shift but with slightly unequal current magnitudes. In this case we need to compensate for that with a calibrated attenuator in the probe line at the hybrid coupler input. It is essential that the probe coaxial cables are terminated in their characteristic impedances so that line length equals phase shift.
- Assume the array is not fed in 90° current increments, but with a phase difference of 111° (such as between the center elements and the back element in the WA3FET Four-Square). All we need to do in that case is insert an additional line length of (111–90) = 21° in the line going

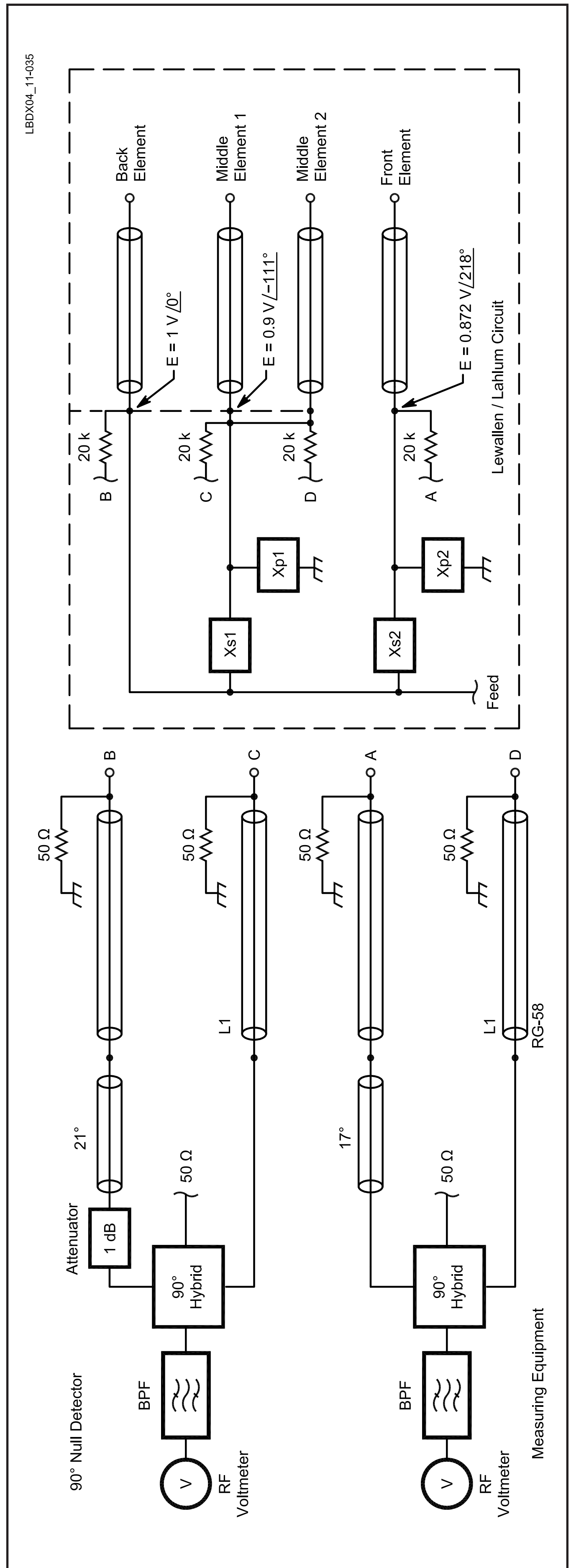



**Fig 11-35—Some RF is sampled at the end of the λ/4 lines going to the antenna elements. This is fed via RG-58 voltage sampling lines of equal length to the measuring equipment. Short line lengths and small attenuators can be inserted to compensate for non-quadrature setups and unequal drive currents. The schematic of the 90° hybrid is given in Fig 11-21. Section 3.4.6 explains how to calculate Xs1, Xp1, Xs2 and Xp2. V is a detector, which can be the detector/wattmeter described in Section 3.6.1 or a receiver. BPF is a bandpass filter.**

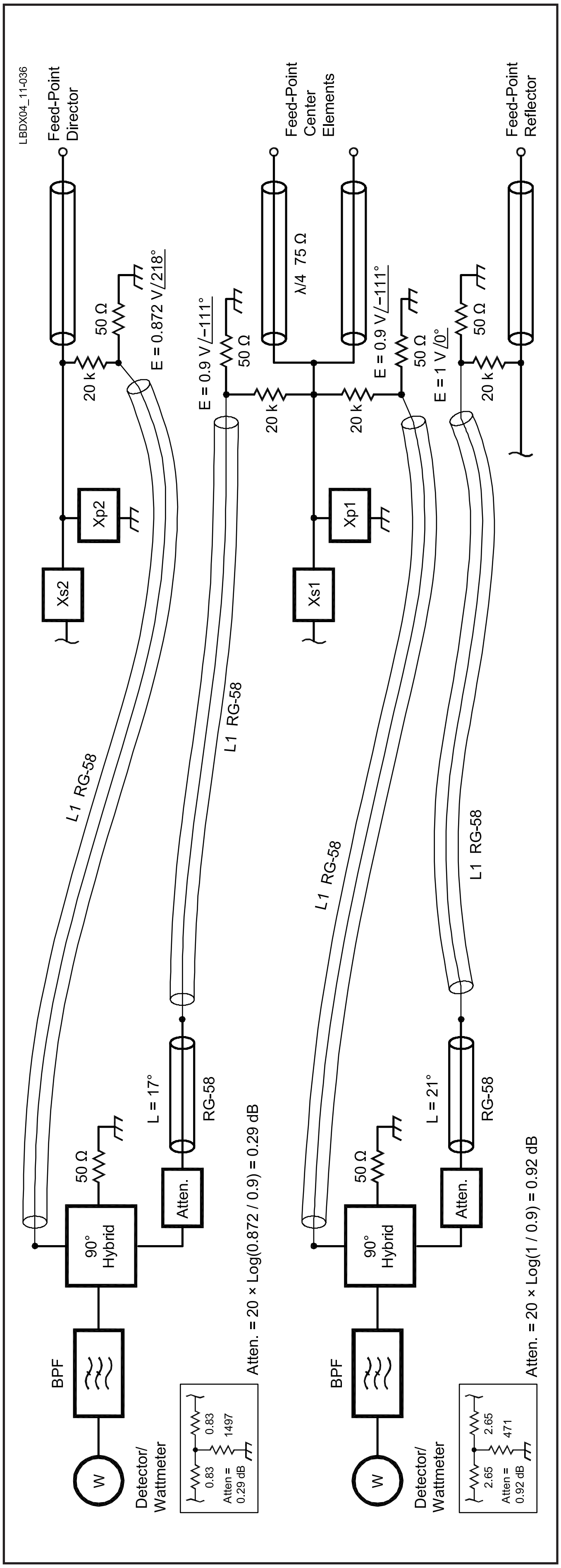



**Fig. 11-36—Detailed schematic of the test setup for the WA3FET optimized Four-Square.**

to the element with the leading phase, so that the net result again is 90°. See **Fig 11-36**.

In the same example the phase difference between the center elements and the director is -107°, hence we need an additional line length in the measuring set up of 17°. When measuring between points A and C, we need to insert a 1-dB (a 0.89:1 voltage ratio) attenuator in the line to point B to compensate for the unequal drive currents. The value of the sampling resistors depends on the power you want to do the testing with, and the detector's sensitivity.

### *3.6.2.1 Discussion on required signal levels, BC interference, and detector sensitivity.*

Ideally we would want to be able to do some testing with an antenna analyzer (eg, the MFJ-259B ) as a signal source, and using a small Detector/Wattmeter as described in Section 3.6.1. This way we can work on the antenna with really portable equipment. This should do for initial tuning even if you are not able to get a null better than 30 dB. As a final touch up, you can always use the station transmitter as a signal source for doing final alignment.

- What are the limiting factors?
- BC signals or even broadband noise.
- Detector sensitivity (noise figure)
- Available testing power

W1MK says that when he starts a measurement session, he first measures the level of background signals or noise on the antenna. For that you simply connect the detector/wattmeter to the antenna you will be testing. A broadband noise level of −35 dBm for 80 meters and even more on 160 is not uncommon, and in some case can be much higher (10 or 20 dB higher!). These values will of course be different in different locations.

Adding a band-pass filter (BPF) in front of the broadband detector should drop the meter readings signficantly. The values, of course, will be different for different locations. For example, W1KM experiences very high levels (–45 dBm) even with a BPF in front of the detector due to strong BC interference levels. In most situations the majority of the power hitting the detector is from out-of-band signals and if not filtered out by a selective circuit will reduce the amount of null that can be obtained. If the interference is inside the BPF, you can apply more power, or use a receiver to provide more selectivity.

For minimum measurement error a sampling resistor value of 20 kΩ is recommended. This means that the sampled signal will be approx −52 dB down from the applied power. If we apply power with the MFJ-259, the level will be +13−52 = approximately –40 dBm.

If we use the detector/wattmeter described in Section 3.6.1 (which has a maximum sensitivity of −75 dBm) and if we are not limited by BC signals, we can see a null down as far as −35 dB. This is not bad for a starter! An S9+40 signal represents −32 dBm, which means that the sensitivity of the detector/wattmeter matches pretty well with the level of a S9+40 signal, and even with such strong broadcast signals you will be able to see nulls of approx –30 to –35 dB.

In case of very stubborn noise/interference problems you can, of course, use your receiver as a null-detector. It has surplus sensitivity and should have enough selectivity to reject offending signals.

Your ability to obtain a deep null with a simple detector/wattmeter will always be either noise limitation (the internal noise or the noise figure of the detector/wattmeter) or interference limitation. If it is out-of-band interference, a BPF will help. If the interference is *on* your desired testing frequency you can move the test frequency slightly, or even better apply more power.

You might use 10-kΩ sampling resistors, if sensitivity is a problem but that is the limit—It is better to use higher testing power. A simple testing procedure is the following:

- Always use a bandpass filter at the input of the detector/wattmeter.
- Start you session with a portable source, such as the MFJ-259 antenna analyzer.
- Adjust the L-network values for maximum null. You should be able to obtain a null of at least –30 dB.
- If you are satisfied with a 30-dB null, now use your exciter as a signal source and apply 10 Watt (+40 dBm). This about 27 dB better than the MFJ-259, which means that under the same circumstances you now will be able to see a null down to 50 dB.

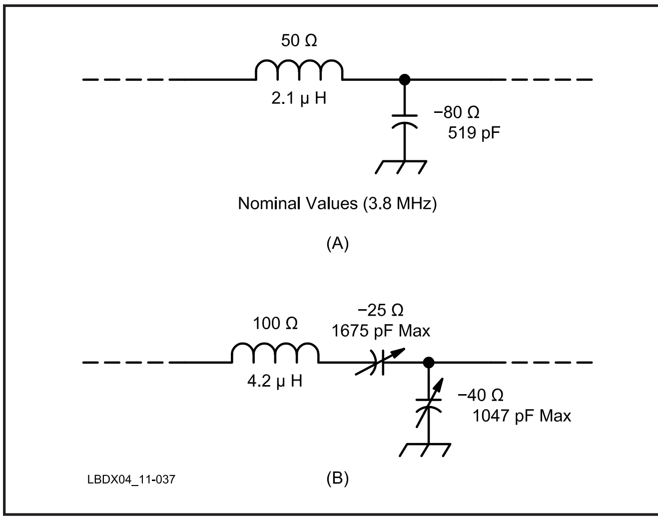



**Fig 11-37—To make the Lewallen L-network continuously adjustable, replace the coil with a coil of twice the required value and connect a capacitor in series. The net result will be a continuously variable reactance. With the values shown, the nominal +50-Ω reactance is adjustable from +75 to +25 Ω (and less). The two capacitors can be motor driven to make the phase-shift network remotely controllable.**

For fine trimming the phase and amplitude you must be able to fine adjust both the series and the parallel reactances of the L-network. A variable capacitor is an obvious choice for fine trimming. You can make the equivalent of a variable inductor with a little trick. For example, if the networks requires a coil with a reactance of +50 Ω, make a coil with double the reactance (100 Ω or 4.2 µH at 3.8 MHz) and connect in series a variable capacitor with (at maximum capacitance) a reactance of –50 Ω or less. If you use –25 Ω (1675 pF at 3.8 MHz), the series connection of the two elements will now yield a continuously variable reactance (at 3.8 MHz) of +25 (or less) to +75 Ω. See **Fig 11-37**.

The nice feature of such a test setup is that you can leave it permanently connected. Make sure that your sampling resistors are of high wattage if you run high power. Using 20-kΩ sampling resistors and running 1500 W the resistors dissipate 3.75 W, so two 40-kΩ, 2-W resistors in parallel is adequate.

The sampled power level going into the hybrid is –50 to –60 dB down from the transmit power, which puts it in the 1 to 10-mW (0 to +10 dBm) level for 1000 W (= +60 dBm) transmit power. A 40-dB null would show up as –30 to –40 dBm on your detector/wattmeter in the shack.

A –30 dBm level is 7 mV in 50-Ω. If you just want a kind of alarm system that tells you when things are really wrong, a simple germanium diode detector and a sensitive analog microamp meter (eg, 50 µA full scale) could be used.

Don't expect to have enough nulling sensitivity with this setup to properly adjust the L-network components. For that you need the sensitive wattmeter in Fig 11-30. To avoid overdriving the detector-wattmeter you should provide a 10/20/30-dB step attenuator when running high power.

### 3.6.3. Measuring antenna resonance

The true resonant frequency is the frequency where the reactive part of the impedance equals zero. You can use one of the common antenna analyzers (see Section 3.6.8.), but their accuracy is not always the best, at least not when compared to the method described below.

W1MK uses the detector/power meter described in Section 3.6.1, together with a so-called *6-dB hybrid* to measure the resonant frequency, as well as the $R_{rad}$ + $R_{loss}$ of the antenna very accurately. The circuit is very simple. See Fig 11-37. It boils down to a resistive bridge, where the detector has an asymmetric input is fed via a balun. The circuit is similar to the old "Antennascope" described 50 years ago in many handbooks, except that W1MK now uses a very sensitive null detector. This allows him to achieve a very deep null and to determine the exact resonant frequency. The signal source is not very critical and a typical antenna analyzer such as MFJ-259B should do.

If you cannot achieve a very deep null, BC band signals or the harmonic content of the signal generator may be a problem. Insert a band-bass filter between the generator and the bridge. The W3NQN bandpass filters are the best in this application. I use them between the exciter and my amplifier, so they are always available for such an application). The 50-Ω, non-inductive resistors must be matched if you want to read the value of the antenna total resistance from the potentiometer scale. T1 is a little balun that can be wound on a FairRite Products 2873000202 core (or similar). Use twisted-pair enameled wire (#24 to #26) to wind six passes (= 3 turns, = 3 times through both holes) on the binocular core.

Robye, W1MK, points out that he made provisions allowing him to actually measure the value of the variable resistor, using his digital multimeter, which allows him to get very accurate results.

Connect the antenna to the ANT terminal and adjust the frequency of the generator and the value of the potentiometer until the deepest null is reached. This will be at the antenna's resonant frequency. The value that you read off the potentiometer is the sum of $R_{rad}$ and $R_{loss}$ of the antenna.

### 3.6.4. Measuring antenna impedance using the W1MK 6-dB hybrid

Although the 6-dB hybrid (or Antennascope) described in Section 3.6.4 is merely is a resistive bridge circuit that can only be nulled when terminated in a purely resistive load, we can still use it to make accurate impedance measurements.

What we need to do is tune out the reactive part of the antenna impedance before it is connected to the bridge. We can do this simply by connecting a coil or a capacitor of the appropriate value in series. (Alternatively you could put the reactance in series with the 100-Ω potentiometer). Once this is done you can read the real part of the antenna impedance from the calibrated potentiometer scale on the 6-dB hybrid. See **Fig 11-38**.

The imaginary part of the impedance is the conjugate value (just change the sign) of the value of the series coil or capacitor used to tune out the reactance. **Fig 11-39** shows the schematic for a unit I built around a beautiful 5 × 4000 pF BC variable with built-in 91:1 gear reduction. In combination with a Groth turns counter, it is possible (after calibration against a laboratory grade instrument) to read off the capacitance over the entire range with an accuracy of a few pF!

With S1 in position a, you can obtain C values from about 100 pF to 6000 pF, which means capacitances ranging from −5.5 Ω to > −500 Ω on 80 meters and −11 Ω to > −1000 Ω on 160 meters. Of course S2 or S2 and S3 will need to be closed for the lower values. If needed, we can always add extra capacitors to obtain even lower values.

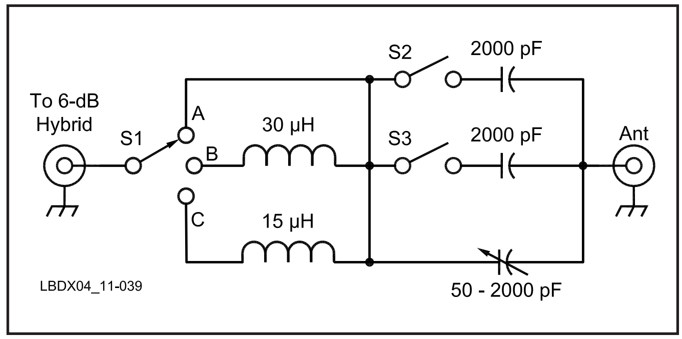



**Fig 11-39—You can always change a complex feed-point impedance of an antenna to a pure resistive impedance (which means bringing it to resonance) by adding the appropriate value of reactance in series. This simple circuit allows you add a wide range of positive, as well as negative, reactances to do this. See text for details.**

With S1 in position b (S1 and S3 open), you can obtain reactance values going from a few ohms to −170 Ω on 160 meters and up to −340 Ω on 80 meters. If that is not enough, we can put S1 in position C, where these values are doubled.

Ideally, this unit should be calibrated using a professional-grade network analyzer or impedance bridge. Once this is done you are all set with a very accurate impedance measurement set-up for antennas.

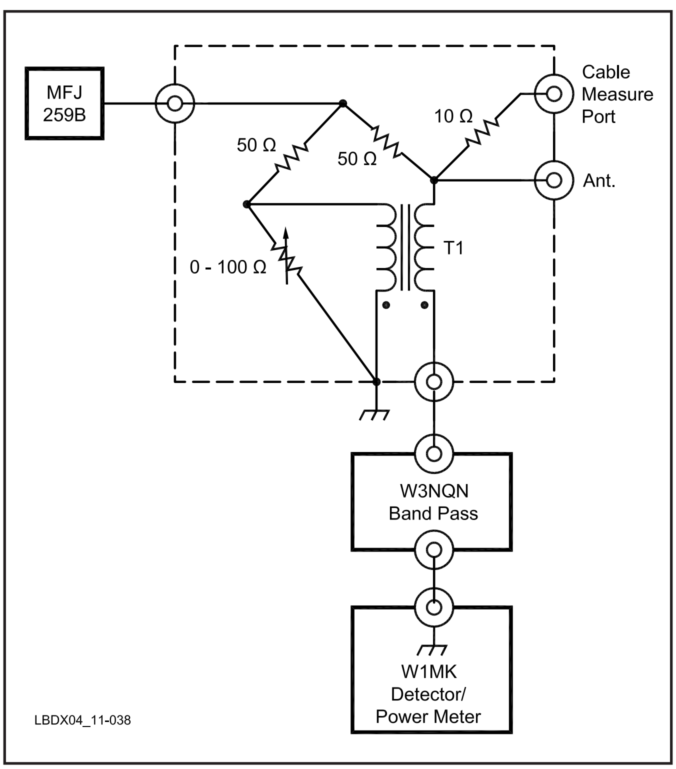



**Fig 11-38—The 6-dB hybrid is the heart of the measuring setup for determining antenna resonance. See text for details.**

### 3.6.5. Using the 6-dB hybrid to make λ/4 lines, λ/2 lines or multiples thereof.

The 6-dB hybrid circuit described in Section 3.6.2 makes an ideal piece of test equipment for cutting stubs. Refer again to Fig 11-38. You can trim λ/4 long lines by leaving the far end open-circuited. For trimming λ/2 lines you can do it with the far end open-circuited on a frequency that is twice the design frequency. You can, of course, also use shorted (at the far end) lines, but make sure the short is a zero inductance short! It is easier to make a perfect open-circuit than a perfect short-circuit. Here is the procedure:

1. First short the "CABLE MEAS PORT" connector, preferably with a coaxial short (not just a wire loop, since that is *not* a very good short at RF).
2. Adjust the generator (antenna analyzer) to the desired frequency, where the feed line will be a short.
3. Next adjust the potentiometer for maximum notch (minimum power as detected by the W1MK detector/power meter). The value should be approximately 10 Ω.
4. Connect the stub, whose far end has been shorted for λ/2 or open-circuited for λ/4, to the "CABLE MEAS PORT" connector.
5. Tune the generator frequency and find the frequency of deepest null while slightly changing the value of the potentiometer for the best null.
6. I hope you started with a stub that was too long! Now cut off short lengths at a time, taking care to preserve a good dead short at the end with no inductance for λ/2, until you are right on the dot.

### *3.6.5.1. Measuring cable loss with this set up*

A lossless λ/4-long cable open-circuited at its end represents a dead short at the other end. The low resistance valued measured is a measure of the loss of the cable. If the cable were truly lossless, the potentiometer setting would be 10 Ω, the value of the series resistor going to the "CABLE MEAS PORT." If the potentiometer reads (10 + a) Ω for bridge balance at the resonant frequency, the cable loss = 8.69 × a × $Z_0$ where $Z_0$ = characteristic impedance of the cable. You must very accurately measure both the 10 Ω resistance and the value of the potentiometer!

Example: $Z_0$ = 50 Ω, a = 1 Ω (that is, the potentiometer is 11 Ω). The Matched Loss (dB) = 8.69 × 1/50 = 0.17 dB. The results are very accurate for a loss up to 5 dB. See **Table 11-7**.

### 3.6.6. Noise bridges

Commercially available noise bridges will almost certainly not give the required degree of accuracy, since rather small deviations in resistance and reactance must be accurately recorded. A genuine impedance bridge is more suitable. But with care, a well-constructed and carefully calibrated noise bridge may be used.

Several excellent articles covering noise bridge design and construction have been published, written by Hubbs, W6BXI; Doting, W6NKU (Ref 1607); Gehrke, K2BT (Ref 1610); and J. Grebenkemper, KI6WX (Ref 1623); D. DeMaw, W1FB (Ref 1620); and J. Belrose, VE2CV (Ref 1621). These articles are recommended reading material for anyone considering using a noise bridge in array design and measurement work.

The software module RC/RL TRANSFORMATION part of the NEW LOW BAND SOFTWARE is very handy for transforming the value of the noise-bridge capacitor, connected in parallel with either the variable resistor or the unknown impedance, first to a parallel reactance value and then to an equivalent reactance value for a series-LC circuit. This enables the immediate computation of the real and imaginary parts of the series impedance equivalent, expressed in "A + *j* B" form. Noise bridges are frequently used to cut quarter-wave or half wave transmission lines:

### *3.6.6.1. Using the noise bridge as a noise source only*

If you have a noise bridge such as the Palomar bridge, you can use it as a wide-band noise source, without using the internal bridge. Instead you will connect the line to be trimmed across the output of the noise bridge and trim the length until the noise level on the receiver is reduced to zero. Switch off the receiver AGC to make the final adjustments (see Fig 11-28B). Tune the receiver back and forth across the frequency to determine the frequency of maximum rejection quite accurately. In this method λ/4 lines should be open-circuited at the end, and λ/2 lines should be short circuited.

### *3.6.6.2. The K4PI method.*

Another method consists in using the noise bridge not only as a noise generator, but also as a bridge. Here is the procedure Mike Greenway, K4PI, uses with great success:

First put a really good RF short at the "UNKNOWN" terminals of the bridge. Using the XL/XC control and the RESISTANCE knobs alternatively, null the noise in the receiver. This is an important step. Keep increasing RF and AF gain and moving the bridge controls to obtain the lowest noise hiss you can. If you do it correctly you will get to the point where the receiver will sound almost dead.

Now, treat λ/4 wave sections as a λ/2 section because the λ/4 method shows too broad a reading. Prepare the short at the end of the coax by removing some of the outer plastic sheath. Push back some of the shield and remove some center-conductor insulation. Pull the shield back and squeeze it onto the center conductor and apply some solder. This makes a good RF short.

Switch the receiver (detector) to AM with the AGC off. The receiver must then be tuned to the area you are expecting to find the null. Connect the coax to the ″UNKNOWN″ terminals taking care not to touch the XL/XC and RX settings. Now use the RESISTANCE knob to null the noise along with tuning the receiver up and down the band for the lowest noise point. If you do everything right and listen very carefully you can get a null on an 80-meter λ/4 line being checked around 7300 kHz (where the stub is λ/2 long) to within 5 - 8 kHz. Take the center of that spread as the true null frequency.

Here too, if you have problems getting a deep null, you may want to try a bandpass filter (eg, W3NQN) between the noise bridge and the receiver. See **Figure 11-40**.

### 3.6.7. Network analyzers

Professional network analyzers are, in principle, ideal tools for measuring impedances. There are various types on the market, and second hand you may be able to get a system with an analyzer and generator for between $1000 and $2000. When measuring antennas on 80 and 160 meters, it is important that you do these measurements during day time, because during the night the average signal power on the band is so great that this background noise will cause erroneous readings on the equipment. With good quality equipment, one can adjust the generator power level to overcome this problem to a certain degree.

### 3.6.8. Measuring antenna impedance using one of the popular antenna analyzers

### *3.6.8.1. The AEA CIA-HF antenna analyzer*

The AEA-CIA-HF analyzer is a one-port network analyzer with limited capabilities. It measures impedances (and of course SWR) by a swept-frequency method, over a range

**Table 11-7**
**Conversion for 50 and 75-Ω systems**

| *a* (Ω) | *Loss (dB) for $Z_0$ =50 Ω* | *Loss (dB) for $Z_0$ = 75 Ω* |
|---|---|---|
| 1 | 0.17 | 0.11 |
| 2 | 0.34 | 0.22 |
| 3 | 0.52 | 0.34 |
| 4 | 0.69 | 0.46 |
| 5 | 0.87 | 0.58 |
| 6 | 1.02 | 0.70 |
| 7 | 1.22 | 0.81 |
| 8 | 1.39 | 0.93 |
| 9 | 1.56 | 1.04 |
| 10 | 1.74 | 1.16 |

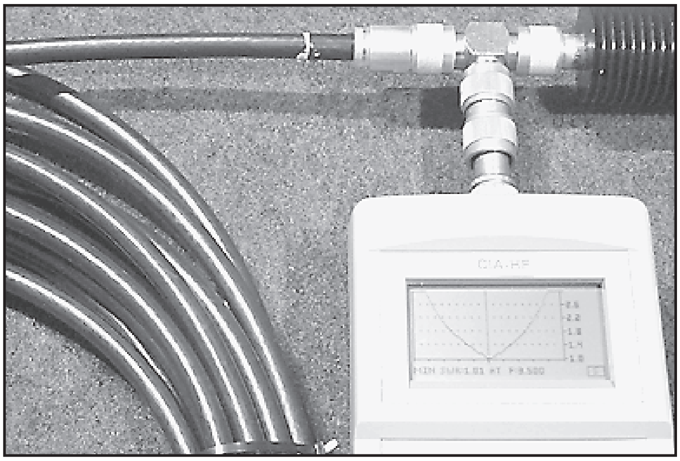


Fig 11-40—The AEA CIA-HF showing the SWR curve and the stub frequency (frequency of minimum SWR). The stub was initially cut for exactly 3.5 MHz using and R&S network analyzer. The text under the graph reads "MIN SWR 1.01 at 3.500 MHz."

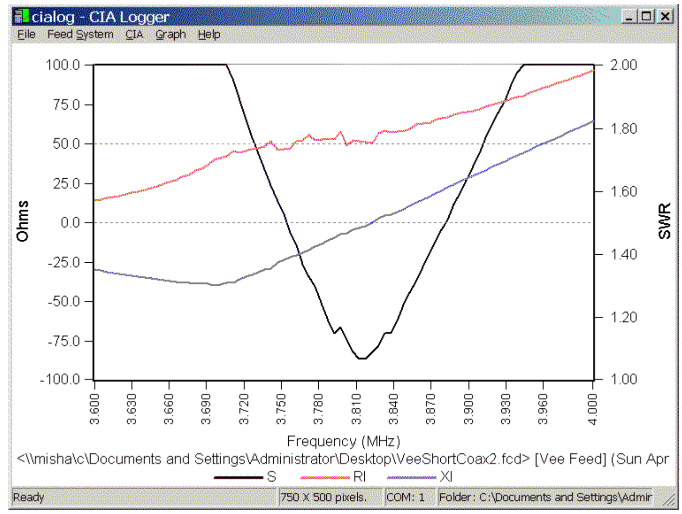



Fig 11-41—Screen shot of the software developed by W8VWW for the CIA-HF analyzer. See text for details.

that you can set between 0.4 and 54 MHz. The nice thing is that it is portable, and can operate from built-in batteries. However the power consumption is pretty high and it's a good idea to run it from a small 12-V power supply in the house or from a small lead-battery on a shoulder strap when in the field. When using it to measure antenna impedances, I have found it quite useful on all bands, down to 40 meters, and sometimes 80 meters. See **Fig 11-40**. On 160 meters, signals picked up from the broadcast band are too strong and mess up the readings, even during daytime.

The challenge will be for someone to come up with filters that will eliminate the BC interference, without causing any impedance transformation in the measuring range. This is quite a challenge. Another solution would be to have a higher output, but that may conflict with the FCC regulations on this subject.

The little screen on the unit does not show much detail, and when you use it in the shack the use a PC with the appropriate control and display software is recommended. AEA (**www.aea-wireless.com/cia.htm**) has such software, called "Via Director". The VIA HF is similar to the CIA-HF but has slightly extended frequency range.

Greg, W8WWV also developed similar software for the CIA-HF, It can be downloaded from his website at **www.seed-solutions.com/gregordy/Software/cialog.htm**. This web page describes the software, and near the bottom there is a link to download the self-extracting program that installs the software. See **Fig 11-41** for a screen shot of the graph produced by W8WWV's software.

AEA now also has an improved version of the CIA-HF, called the VIA-Bravo, which goes all the way up to 200 MHz. The VIA-Bravo provides greater accuracy in all complex measurements including 0.01° phase-angle resolution at lower angles. The unit, however, is very expensive.

### *3.6.8.2. The MFJ-259B antenna analyzer*

The MFJ-259B antenna analyzer is different from the older MFJ-259. It uses a microprocessor and four voltage detectors in a bridge to directly measure reactance, resistance and VSWR. With so much information available, uses are limited mostly by your imagination and technical knowledge.

The main application for antenna builders is its capability of measuring SWR (also in terms of reflection loss). It will also measure the resistive part and the absolute value of the reactive part of a complex impedance. The MFJ-259B isn't smart enough give the sign of the reactive part without some minor help. You must vary the frequency slightly and watch the reactance change to determine the sign of the reactance and the type of component required to resonate the system. If adjusting the frequency slightly higher increases reactance (X), the load is inductive and requires a series capacitance for resonance. If increasing frequency slightly reduces reactance, the load is capacitive and requires a series inductance for resonance. This general rule works with most antennas, but not necessarily all of them.

The designers of the unit have added a "transparent filter" to cope with the problems of strong signals messing up low-level reflected-power readings in the vicinity of broadcast transmitters or during night time measurements on the low bands. This accessory includes an adjustable notch filter and selective bandpass filter. This handy accessory allows the MFJ-259B to be used on large low band antennas, even if the antenna is located in the area of a broadcast transmitter.

At first blush the major difference between the MFJ unit and the AEA unit is the fact that the MFJ-unit does not generate a spectral display of the units measured. It is basically a single-point (one frequency) measurement system.

### *3.6.8.3. Cutting stubs with antenna analyzers*

All of the popular antenna analyzers can be used in this application. The method consists of connecting a 50-Ω dummy load to the analyzer via a T connector. The transmission line or stub is connected in parallel with the dummy load. The antenna analyzer is then adjusted for the frequency with the lowest SWR ratio. For an open-circuited cable this is at the frequencies where the cable is λ/2 or multiple thereof. For a short-circuited stub this is for a length of λ/4 or any odd multiple thereof. The AEA CIA-HF Analyzer has a nice feature where it can calculate the frequency of lowest SWR and show it on the screen.

Using the AEA CIA-HF Analyzer, it is possible to determine very accurately the frequency of minimum SWR (which equals the frequency where the stub is resonant). When measuring a stub that was cut for 3.5 MHz using the R&S network analyzer, the average of a number of measurements gave 3.492 MHz for the stub-resonant frequency, which agrees within 0.2%, an excellent figure.

#### *3.6.8.4. Which antenna analyzer?*

On the subject of analyzer measurement accuracy, W8WWV did some elaborate testing comparing some of the popular units . The detailed information is available on his site: (**www.seed-solutions.com/gregordy/Amateur%20Radio/Experimentation/EvalAnalyzers.htm**).

### 3.6.9. N2PK VNA (Vector Network Analyzer)

Genuine network analyzers are expensive, even second hand, but they provide much better accuracy than the present day Antenna Analyzers. At the time this Fourth Edition goes to press, it appears that a Vector Network analyzer developed by N2PK will be a valid replacement for these expensive instruments, and provide results with comparable accuracy. See **Fig 11-42**.

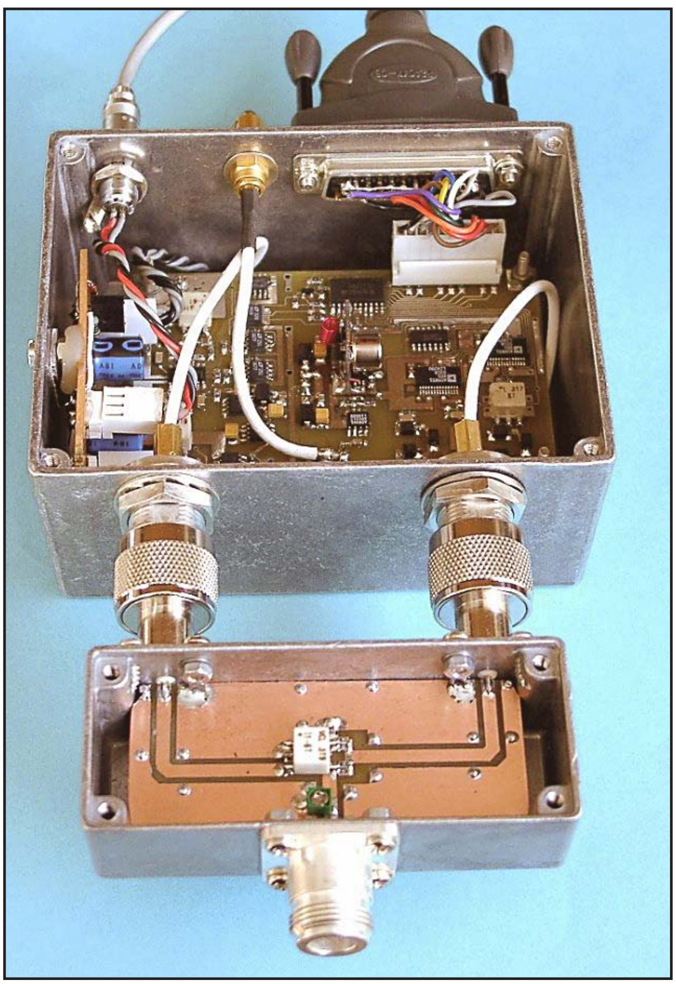


**Fig 11-42—The N2PK-designed VNA (Vector Network Analyzer) constructed by G3SEK. This is all the hardware needed! A PC does the control and interface, of course. It is very likely that soon a commercial version will be available.**

This unit is capable of both transmission and reflection measurements from 0.05 to 60 MHz, with about 0.035-Hz frequency resolution and over 110 dB of dynamic range. Its transmission-measurement capabilities include gain/loss magnitude, phase and group delay. Its reflection-measurement capabilities include complex impedance and admittance, complex reflection coefficient, VSWR and return loss.

Unlike other impedance measuring instruments that infer the sign of the reactance (sometimes incorrectly) from impedance trends with frequency, a VNA is able to make this determination from data at a single frequency. This is a direct result of measuring the phase as well as the magnitude of an RF signal at each test frequency.

N2PK (**users.adelphia.net/~n2pk/**) impressed all of us hams looking for an affordable network analyzer by the level of documentation that he has made available for anyone wanting to build a unit. And the performance is *much* better than anything else you might be able to build or buy for the amount of money you will spend building his VNA. However, building a VNA is not for a first-time kit builder, although there are interest groups supporting potential builders (**www.seed-solutions.com/gregordy/Amateur%20Radio/Experimentation/N2PKVNA/N2PKVNA.htm**).

Even better news is that a "plug and play" commercial version may be available soon. The basic VNA unit works all the way up to 60 MHz, so it's got plenty of range for the low-band enthusiast. Comparing the measurements of the VNA with a top grade network analyzer shows that it is *very* close to a professional instrument.

### 3.6.10. The good old impedance/admittance bridge

All the above-mentioned instruments have the same intrinsic problem of suffering from alien-signal overload when measuring large antennas on the low bands, especially 160 meters, where BC signals are likely to cause false readings unless clever computer algorithms are used to compensate for them. The only other way to overcome this problem is to measure with more generator power, which to a degree is possible with professional-grade network analyzers. Or in the worst of cases, you may have to resort to a good old General Radio bridge, driven by a signal source of sufficient level. This method, of course, lacks the flexibility of a real frequency-sweeping network analyzer.

## 3.7. Mutual-Coupling Issues

### 3.7.1. Too little mutual coupling where you want it

When we set up an array, we need to calculate the mutual impedance from the measurements of the self impedance and the coupled impedance (see Section 3.3.1).The normal procedure is to first measure the self impedance, and then couple one element at a time and measure the coupled impedance.

If you measure little or no difference between the self impedance and the coupled impedance, then have a look at the value of the self impedance. It is likely that the resistive part of the impedance is much higher than the impedance you have calculated by modeling the antenna. For example, if you use inverted-L elements with λ/8 vertical portions, you should expect a self impedance of approximately 17 Ω over a perfect ground. If you measure 50 Ω, it means that you have an equivalent loss resistance of 33 Ω! With so much loss resistance you will see—even with very close coupling such as an array with λ/8 spacing—only a little difference between self impedance and coupled impedance. Such an array will still show the proper directivity, but its gain will be way down. In the above example the gain will be down 4 to 5 dB from what it would be over an excellent ground system. So, if you see no effect of mutual coupling where you *should* see it, suspect you have large losses involved somewhere.

### 3.7.2. Unwanted mutual coupling.

There are cases where you don't want to see the effect of mutual coupling. But they are there and you want to control them. If you happen to have towers (or other metal structures or antennas) within λ/4 of one of the elements of an array, you may induce a lot of current into that tower by mutual coupling. The tower acts as a parasitic element, which will upset the radiation pattern of the array and also change the feed impedances of the elements and the array. To eliminate the unwanted effect from the parasitic coupling proceed as follows:

- Decouple all the elements of the array, with the exception of the element closest to the suspect parasitic tower. For quarter-wave element decoupling, this means lifting the elements from ground.
- Measure the feed-point impedance of the vertical under investigation.
- If a suspect tower is heavily coupled to one of the elements of the array, a substantial current will flow in it. Probe the current by one of the methods described by D. DeMaw, W1FB (Ref *W1FB's Antenna Notebook*, ARRL publication, 1987, p 121) and shown in **Fig 11-43**. If there is an appreciable current, you will have to *detune* the tower. Methods for detuning a tower are given in detail in Chapter 7 (Section 3.7.2).
- After detuning the offending tower, measure the feed-point impedance of the vertical again. If you have properly detuned the parasitic tower, you will likely see a rise in impedance and a shift in resonant frequency.
- Reconnect the whole array and fire in the direction of the parasitic tower.
- Check the current in the parasitic tower and if necessary make final adjustments to minimize the current in the tower. You can use high power now to be able to tune the tower very sharply. In general the tuning will be quite broad, however.
- You now have made the offending tower invisible to your array.

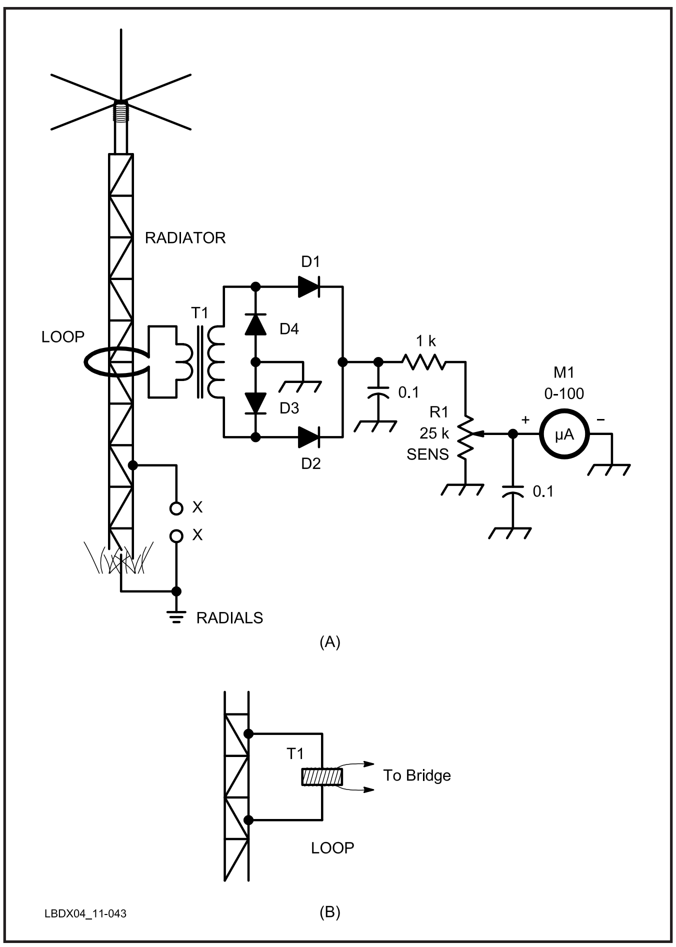



**Fig 11-43—Current–sampling methods for use with vertical antennas, as described by DeMaw, W1FB. Method A requires a single-turn loop of insulated wire around the tower. The loop is connected to a broadband transformer, T1. A high-mu ferrite toroid, as used with Beverage receiving antennas (see Chapter 7 on special receiving antennas), can be used with a 2-turn primary and 2 to 10-turn secondary, depending on the power level used for testing.**

## 3.8. Network Component Dimensioning

When designing array feed networks using the computer modules from the NEW LOW BAND SOFTWARE, you can use absolute currents instead of relative currents. The feed currents for the 2-element cardioid array (used so far as a design example) have so far been specified as $I1 = 1 \angle -90°$ A and $I2 = 1 \angle 0°$ A. The feed-point impedances of the array are:

$$Z1 = 51 + j\,20\ \Omega$$
$$Z2 = 21 - j\,20\ \Omega$$

With 1 A antenna current in each element, the total power taken by the array is 51 + 21 = 72 W. If the power is 1500 W, the true current in each of the elements will be:

$$I = \sqrt{\frac{1500}{72}} = 4.56\ \text{A}$$

Using this current magnitude in the relevant computer program module COAXIAL TRANSFORMER will now show the user the real current and voltage information all through the network design phase. The components can be chosen according to the current and voltage information shown.

If you plan to build your own Lahlum/Lewallen network, it's a good idea to stick to air-wound coils (have a look at Fig 11-27) for inductances up to approx 5 µH. Above this value you will have to revert to toroidal cores. Ferrite cores should not be used in this application since they tend to be unstable under certain circumstances. Only use powdered-iron cores. The red cores (mix 2) are a good choice for both 160, 80 and 40 meters. How large a core do you need to use? The rule is never to wind more than a

**Table 11-8**
**Maximum inductance for a single layer winding, as a function of wire diameter**

| *Type* | *$A_L$* | *#10* | *#12* | *#14* | *#16* | *#18* | *#20* |
|---|---|---|---|---|---|---|---|
| **T106-2** | 135 | 3.9 | 6 | 9 | 15 | 22 | 35 |
| **T157-2** | 140 | 12 | 18 | 24 | 47 | 68 | 110 |
| **T200-2** | 120 | 16 | 25 | 40 | 65 | 95 | 153 |
| **T200A-2** | 218 | 29 | 46 | 73 | 119 | 172 | 278 |

single layer. **Table 11-8** gives you the maximum inductance that you can get with a given wire size (AWG #) for a given core.

Example: Assume you need a reactance of +800 Ω. On 1.83 MHz that represents 69 µH. You may marginally make it on a T157-2 core with #18 wire. In most cases where such high values of inductance are involved, current through the coil will be very small and #18 enameled wire would be just fine. Only in cases were inductances of between 10 and 15µH are required, I would use A T200 or T200A core with #10 or even #8 wire.

## 4. POPULAR ARRAYS

Whereas in previous editions I described in detail how various feed systems can be applied to various arrays, I decided to describe mainly two feed systems (with one exception) in detail:

- The hybrid-coupler method (plug and play), where applicable for quadrature feeding
- The Lewallen/Lahlum feed method, which allows the most flexibility.

All arrays were modeled using *NEC-2* over "good ground"(conductivity 5 mS/m, dielectric constant 13), with an extensive radial system that accounts for an equivalent-series-loss resistance of 2 Ω for each element. The element feed-point impedances shown include this 2 Ω of loss resistance. If you want to calculate your feed system for different equivalent-ground-loss resistances, apply the following procedure:

- Take the values from the array data (see further). The resistive part includes 2 Ω of loss resistance. If you want the feed-point impedance with 10 Ω of loss resistance, just add 8 Ω to the resistive part of the feed-point impedance shown in the array data. The imaginary part of the impedance remains unchanged.
- Follow the feed-system design criteria as shown, but apply the new feed-point impedance values.

I did the modeling using a wire diameter of 200 mm (approximating a Rohn 25 tower) for the vertical element, and the elements were adjusted to resonance, with all other elements decoupled, meaning floating.

The gain is expressed in dBi (over good ground as specified above). For each array we also calculated the directivity, expressed in RDF (Receiving Directivity Factor) and in DMF (Directivity Merit Figure). See Chapter 7.

In many arrays you will see a negative impedance, in most cases for the "back" element of the array. Again, the negative impedance merely means that the feed network is not supplying power to that element but rather taking power from that element. The different modules of the NEW LOW BAND SOFTWARE as well as the *Lahlum.xls* spreadsheet program handle these negative values without problems.

All Lahlum/Lewallen feed networks are calculated without taking into consideration the effects of cable losses. These effects are quite small on the low frequency bands, if good cables are used. Only with very long cable lengths (eg, 3λ/4 current-forcing feed lines plus a 180° phasing line, losses can be significant. I made several calculations between ideal case (no losses) and the real-world case, and the differences of the L-networks values were well within the typical tuning range of the components. When you take into account the losses, the feed impedance of the network will be slightly higher (typically a few percent).

### 4.1. Two-Element End-Fire Arrays

The principles of operation of the 2-element end-fire array were explained in detail in Chapter 7. Most of us probably think of a λ/4 spaced array, where the elements are fed 90° out-of-phase, but this is not necessarily the best solution. If you want to use 90° phase shift, for instance because you want to use a hybrid coupler to feed the array, then a spacing of about 105° achieves just marginally better DMF than 90°. Staying with quarter-wave spacing, a phase difference of

**Table 11-9**
**Main Data for a Range of 2-Element End-Fire Arrays**

| *Spacing (dBi)* | *Phase BW* | *Gain dB* | *3-dB dB* | *RDF ele.* | *DMF ele.* | *Zfront* | *Zback* |
|---|---|---|---|---|---|---|---|
| 105° | –90° | 4.24 | 177 | 8.13 | 13.1 | 55 + *j* 13 | 19 – *j* 13 |
| 90° | –90° | 4.23 | 177 | 8.11 | 12.3 | 53 + *j* 18 | 21 – *j* 19 |
| 90° | –105° | 4.72 | 159 | 8.70 | 14.4 | 48 + *j* 21 | 17 – *j* 14 |
| 90° | –110° | 4,87 | 154 | 8.88 | 15.1 | 46 + *j* 23 | 16 – *j* 15 |
| 75° | –120° | 5.05 | 146 | 9.14 | 15.6 | 37 + *j* 25 | 14 – *j* 15 |
| 60° | –135° | 5.20 | 135 | 9.50 | 17.2 | 24 + *j* 24 | 12 – *j* 15 |
| 45° | –145° | 5.00 | 131 | 9.57 | 16.6 | 14 + *j* 19 | 10 – *j* 17 |

about 105° is recommended, in which case you can no longer feed the array with a hybrid coupler. The larger the array the better the bandwidth, and this shows in the element impedances. Small arrays, such as those λ/8 spacing, give excellent directivity but the element feed impedances become low, causing drop in gain or a given ground system and small bandwidths over which directivity will hold.

### 4.1.1. Data, 2-element end-fire array

**Table 11-9** shows the main data for a range of 2-element end-fire arrays. The first impression is that 60° spacing with 135° phase shift is best, but note the relatively low feed impedance, which means narrower bandwidth than for a wider-spaced array. The gain figures are over average ground ($\varepsilon = 13$ and $\sigma = 5$ mS).

### 4.1.2. Feed systems, 2-element end-fire array

Several feed methods were illustrated with a 2-element end-fire array in Section 3.4.

#### *4.1.2.1. Christman feed, 2-element end-fire array*

See Section 3.4.2. This approach uses a minimum of components, but since it does not use current-forcing feed lines you cannot measure voltage to determine the feed current. This means you either need to be able to measure feed current (not so easy to do accurately), or you need to do some precise element-impedance measurements (coupled and uncoupled), calculate the mutual coupling and from there figure the actual feed impedances. (You can use the module MUTUAL IMPEDANCE AND DRIVING IMPEDANCE from the NEW LOW BAND SOFTWARE.)

Fig 11-7 shows how you can switch the array in the two end-fire directions. When both elements are fed in-phase the array will have a bi-directional broadside pattern (see Section 4.2) with a gain of 1 dB over a single vertical. The front-to-side ratio is only 3 dB. The feed impedance of two quarter-wave-spaced elements fed in-phase is approximately $57 - j\,15\ \Omega$, assuming an almost-perfect ground system with 2-Ω equivalent-ground-loss resistance. Notice that both elements have the same impedance, which is logical since they are fed in-phase.

We can easily add the broadside direction (both elements fed in-phase) by adding a switch or relay that shorts the 71° long phasing line, as shown in **Fig 11-44**. L networks can be designed to match the array output impedance to the feed line. Don't forget that you need to measure impedances to calculate the line lengths that will give you the required phase shifts. Merely going by published figures will not get you optimum performance!

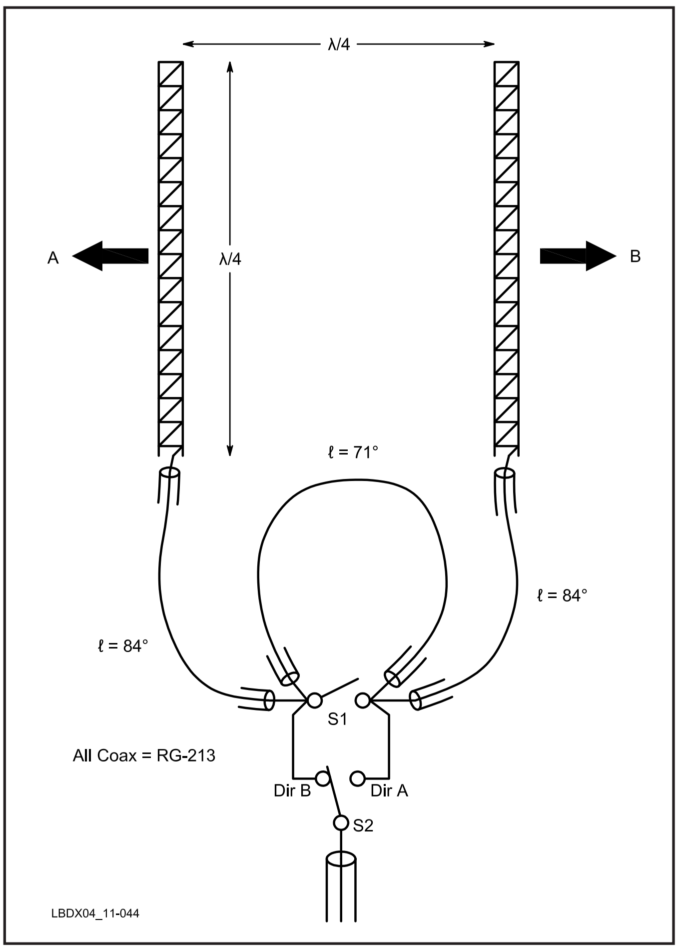



**Fig 11-44—The 2-element vertical array (λ/4 spacing) can be fed in-phase to cover the broadside directions. I added switch S1 to the Christman feed system as described in Fig 11-8. When S1 is closed, both antennas are fed in-phase, resulting in bi-directional broadside radiation.**

#### *4.1.2.2. Hybrid-coupler feed, 2-element end-fire array*

See Section 3.4.6. When you buy a commercial hybrid coupler, you don't really need to do any impedance measurements. All you will have to do is trim the elements to resonance (decoupled from one another!). Commercial hybrid couplers are made to accommodate Four-Square arrays, and normally use four relays to do the direction switching. For a 2-element end-fire array, a much simpler switching system, using a single DPDT relay will do the job if only the two end-fire directions are required.

In this case you can delete K1 and its associated wiring from the schematic shown in **Fig 11-45.** On the low-bands any 10-A relay will do. If you want the bi-directional broadside pattern as well, two relays and an L-C network are needed.

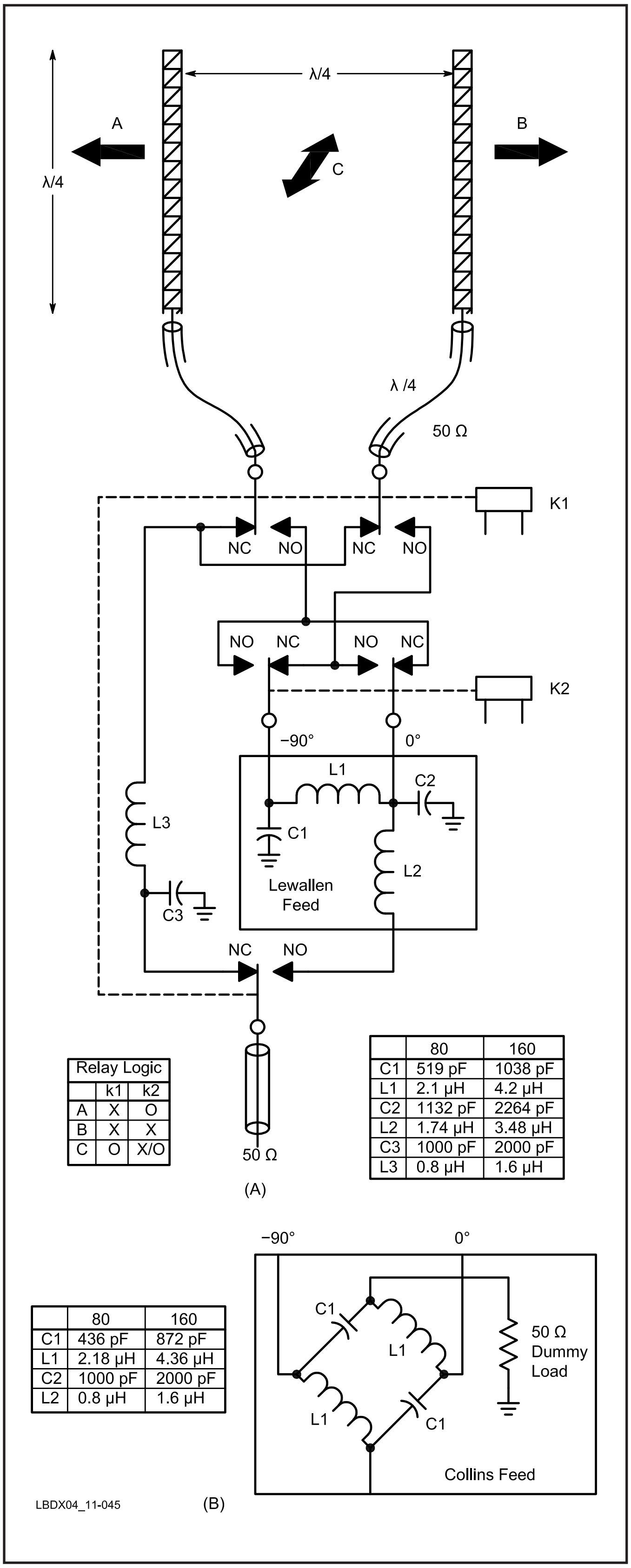
#### *4.1.2.3. Lewallen feed, 2-element end-fire array*

The application of the Lewallen feed method for the 2-element end-fire array was described in detail in Section 3.4.5 and is shown in Fig 11-45. Two-element end-fire arrays are commonly used in a broadside/end-fire combination to increase directivity and gain.

Using the Lewallen feed system, you can adjust the L-network values to obtain the proper feed current magnitude and phase shift, using the simple test method and equipment developed by Robye, W1MK, and described in Section 3.6.

## 4.2. The 2-Element Broadside Array

If you feed two elements in-phase, they will produce a broadside (radiation in a direction perpendicular to the line connecting the two elements) bidirectional figure-eight pattern, provided the spacing is wide enough. The array with 90° spacing is often used as a "third" direction with an end-fire array and gives about 1 dB gain over a single vertical.




| Relay Logic | | |
|---|---|---|
| | k1 | k2 |
| A | X | O |
| B | X | X |
| C | O | X/O |

| | 80 | 160 |
|---|---|---|
| C1 | 519 pF | 1038 pF |
| L1 | 2.1 μH | 4.2 μH |
| C2 | 1132 pF | 2264 pF |
| L2 | 1.74 μH | 3.48 μH |
| C3 | 1000 pF | 2000 pF |
| L3 | 0.8 μH | 1.6 μH |

| | 80 | 160 |
|---|---|---|
| C1 | 436 pF | 872 pF |
| L1 | 2.18 μH | 4.36 μH |
| C2 | 1000 pF | 2000 pF |
| L2 | 0.8 μH | 1.6 μH |

**Fig 11-45— The 2-element vertical array (λ/4 spacing) can be fed in-phase to cover the broadside directions. Two feed methods are shown: At A, the Lewallen feed method, and at B, the Collins hybrid-coupler method. In both these cases relay K1 chooses between the end-fire and the broadside configurations. Relay K2 switches directions in the end-fire position.**

### 4.2.1. Data, 2-element broadside arrays

Narrow spacing yields a wide forward pattern. When we reach λ/2 spacing, and up to about 5λ/8 spacing, the forward lobe is at its narrowest without excessive sidelobes. At λ/2 spacing the rejection off the side is maximum at zero elevation angle. Increasing the spacing lifts the maximum rejection off the ground, resulting in better directivity and higher gain (by way of narrower forward lobe).

See **Table 11-10**. Gain is over average ground, and includes the effect of a 2-Ω equivalent-ground-loss resistance in each element.

### 4.2.2. Feed systems, 2-element broadside arrays

As the elements are fed in-phase, you can feed them with equal-length feed lines to a common point where you parallel the ends of the feed lines. In principle the array can be fed with two feed lines of any equal lengths. Feeding via λ/4 or 3λ/4 feed lines, however, has the advantage of forcing equal currents in both elements, whatever the difference in element impedances might be. I therefore advise people to feed the array via two 3λ/4 feed lines. Quarter-wave feed lines are too short (due to the coax's velocity factor) to reach the center of the array.

Using the COAX TRANSFORMER/SMITH CHART and the PARALLEL IMPEDANCES modules of the NEW LOW BAND SOFTWARE program, you can easily calculate the feed impedance of this antenna. Let's work out an example of a broadside array with 193° spacing:

$Z_{elem} = 28 - j\ 12\ \Omega$

Assume loss-free cables: The impedance at the end of 3λ/4-long current-forcing feed lines ($Z_0$ = 50 Ω) is: 75.4 + *j* 32.3 Ω. Paralleling the two feed lines yields: Z = 38.7 + *j* 6.1 Ω.

Run the SHUNT/SERIES IMPEDANCE NETWORK MODULE and find out that by putting a reactance of −109 Ω (a capacitor) in parallel with this impedance, transforming it into 45 Ω, an almost perfect match for the 50-Ω feed line.

## 4.3. Three- and Four-Element Broadside Arrays

If more than two elements are used in a broadside combination (all in-line and fed in-phase), the current magnitude should taper off towards the outside elements to obtain the best directivity and gain. This current distribution is what is called the *binomial current distribution*. Multi-element broadside arrays are also covered in Chapter 7 on receiving arrays.

**Table 11-10**
**Data for 2-Element Broadside Arrays**

| *Spacing* | *Gain (dBi)* | *3-dB Beamwidth* | *Impedance* |
|---|---|---|---|
| 90° | 2.31 | — | 56 − *j* 17 |
| 135° | 3.52 | 90° | 41 − *j* 19 |
| 180° | 4.91 | 64° | 30 − *j* 14 |
| 193° | 5.26 | 59° | 28 − *j* 12 |
| 208° | 5.59 | 55° | 27 − *j* 9 |
| 225° | 5.81 | 50° | 25 − *j* 5 |

### 4.3.1. Data, 3- and 4-element broadside arrays

The radiation pattern is similar to what is shown in Fig 11-5, only the patterns get narrower and the gain increases as we use more elements. See **Table 11-11**.

### 4.3.2 Feed systems, 3- and 4-element broadside arrays

#### *4.3.2.1. Feed systems, 3-element broadside array*

If we design the array with λ/2 spacing between the elements, the feed lines will need to be 3λ/4 long if we want to follow the current-forcing principle. To obtain double the feed current magnitude in the center element, we need to feed the central element with two parallel feed lines. Using 75-Ω coax for the feed lines we have at the end of those feed lines:

Outer elements: 144 + *j* 182 Ω
Center element: 38 + *j* 17 Ω

Connected in parallel we obtain an array feed impedance of: 27 + *j* 16 Ω, which we can easily match with an L-network to 50 Ω.

#### *4.3.2.1. Feed systems, 4-element broadside array*

Here too, if we want to use current-forcing feed lines, we will need to use 3λ/4 feed lines to the center elements and 5λ/4 feed lines to the outer elements. This involves a lot of coax. If instead of spacing the elements λ/2 we space them 0.8 ×λ/2 (0.8 being the velocity factor of foam coax), we will reach out with λ/4 feed lines to the center elements and 3λ/4 lines to the outer elements. To maintain good directivity and well-suppressed side lobes for this particular case, the current magnitude distribution along the elements is 1:2:2:1. There is some loss in gain vs the λ/2-spaced array (6.8 vs 7.2 dBi),and the 3-dB bandwidth is now 42°. The feed impedances are: 31 − *j* 23 Ω for the outer elements and 36 − *j* 24 Ω for the center elements. To obtain a relatively high total-array feed impedance, it is best to use 75-Ω current-forcing feed lines. We need to run two cables in parallel to the two central elements and single feed lines to the outer elements. The impedance at the end of those feed lines are:

Outer elements: 117 + *j* 87 Ω
Inner elements: 27 + *j* 18 Ω

All connected together, the impedance is 11 + *j* 7.5 Ω. We can match this to a 50-Ω feed line with an L-network.

## 4.4. The 3-Element End-Fire Arrays

We have covered the 2-element end-fire arrays in Section 4.1. Just as we have 2- and 3-element Yagis, we can have 2- and 3-element end-fire arrays.

As we have seen with 2-element end-fire arrays, there is nothing sacred about spacing or phase angles. It is true, of course, that an array with quadrature feeding (phasing angles that are in 90° steps) with identical current magnitudes have

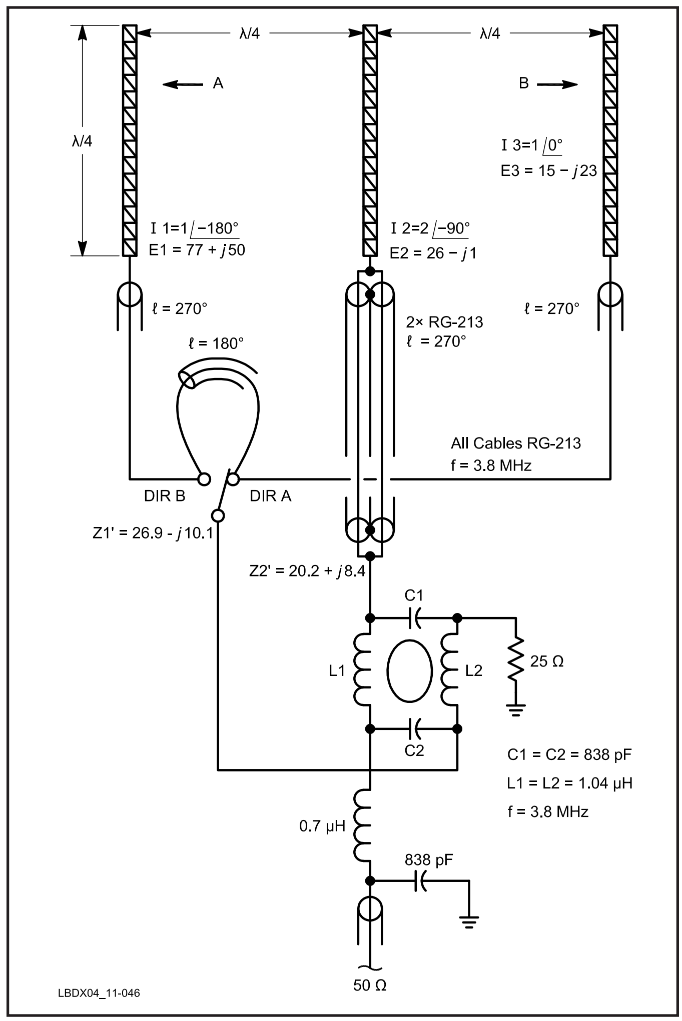



**Fig 11-46—Feed system for the 3-in-line broadside array with binomial current distribution and quadrature phase currents. The center element is fed via two parallel 75-Ω feed lines to obtain double the feed-current magnitude. The current-forcing method ensures that variations in element self-impedances have minimum impact on the performance of the array.**

**Table 11-11**
**Data, 3- and 4-Element Broadside Arrays**

| *Array* | *Element Feed Currents (1)* | *Gain dBi* | *3-dB Beamwidth* | *Element Feed Impedances (Ω)* |
|---|---|---|---|---|
| 3 ele | 1, 2, 1 | 6.33 | 46° | 25 – *j* 19; 31 – *j* 14; 25 – *j* 19 |
| 4 ele | 1, 3, 3, 1 | 7.21 | 37° | 23 – *j* 23; 29 – *j* 16; 29 – *j* 16; 23 – *j* 22 |
| 4 ele (2) | 1, 2, 2, 1 | 6.80 | 42° | 31 – *j* 23; 36 – *j* 24; 36 – *j* 24; 31 – *j* 23 |

(1) Current magnitude
(2) Element spacing = 0.4 λ (see text)

## THE λ/4-SPACED ARRAY—END-FIRE AND BROADSIDE

**Feed-Current Phasing in an End-Fire Array**

For a 2-element array spaced 90° (λ/4), varying the phase of the feed current can be used to not only increase the gain, but also to shift the position of nulls in the rearward direction. **Fig A** shows the physical layout of two λ/4 verticals.

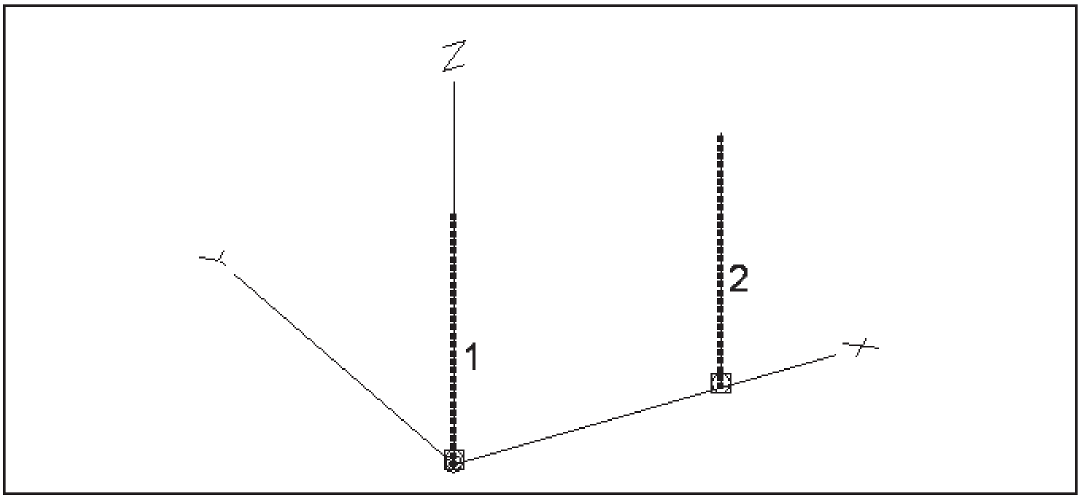



**Fig A—Quarter-wave long vertical elements are positioned on the X-axis. For this example of end-fire operation, element number 1 is fed at 0° phase angle, while element number 2 is fed at either a –90° or a –110° phase angle. Both feed currents have the same magnitude. For broadside operation, both elements are fed with equal-amplitude currents at the same 0° phase angle.**

**Fig B** illustrates how the azimuth patterns change for this array with end-fire feed-current phases of 0° (broadside operation), –90° and –105°. In the 0° phase configuration, the gain decreases compared to the end-fire configurations as the pattern becomes closer to "omnidirectional" but the peak gain is rotated 90° from the peak for the end-fire array—hence the name "broadside."

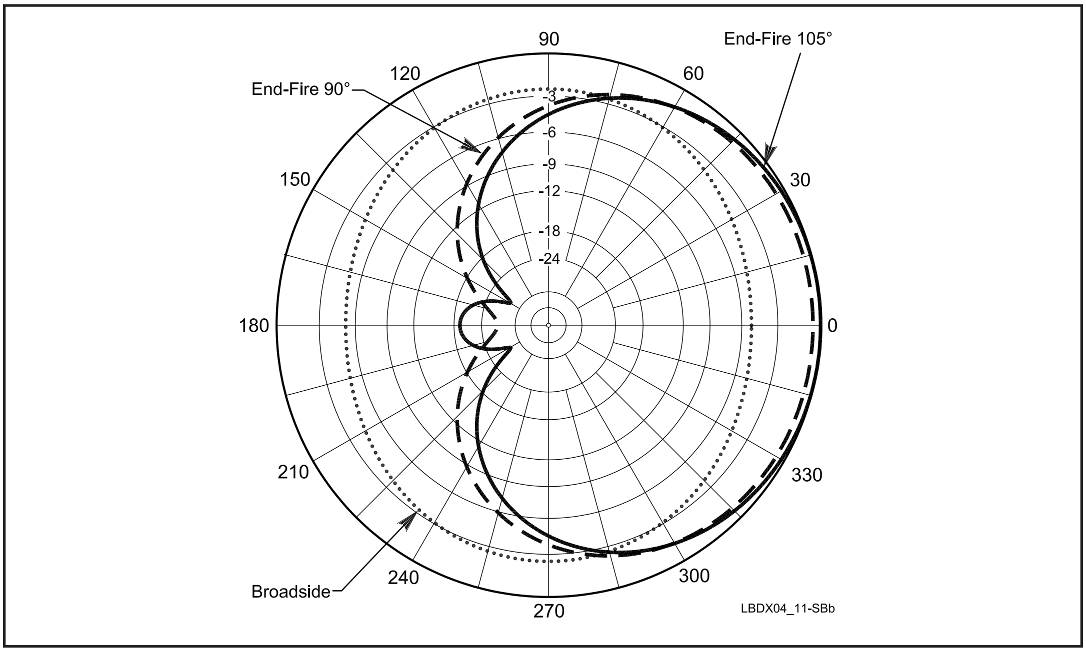



**Fig B—Comparison of azimuthal patterns (at 20° elevation angle) for 2-element vertical array in Fig A, operated end-fire at phases of –90° and –105°. Also shown is response of the array operated at 0° phase, which is the broadside feed configuration. Each element is physically spaced λ/4 from the other. The end-fire peak is along the line between the two elements and is greater than the broadside peak, which is perpendicular to the line between the elements.**

**Fig C** shows the elevation-plane patterns for the two end-fire and one broadside arrangements for the λ/4-spaced 2-element vertical array.

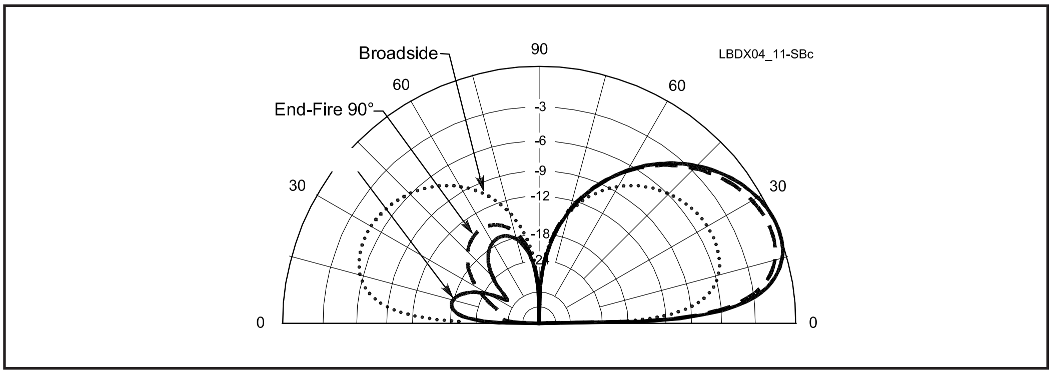



**Fig C—Comparison of elevation-plane patterns for end-fire and broadside arrays shown in Fig A. The 110° phase shift used in the end-fire array puts a null at about a 40° rearward elevation angle and achieves a much better overall directivity in the back compared to a quadrature (90°) phase shift. Of course, the end-fire gain is also higher than the more "omnidirectional" gain of the broadside array.**

### Changing the Element Spacing for Broadside Operation

If the physical spacing between the elements in a 2-element array operated in broadside is varied, the gain will increase with increasing spacing beyond 90° (λ/4). However, more than a spacing of about 225° (5λ/8) results in objectionable sidelobes in the azimuth-plane pattern, as illustrated in **Fig D**. The gain is largest and the sidelobe pattern is cleanest at 193° (0.536 λ) physical spacing.

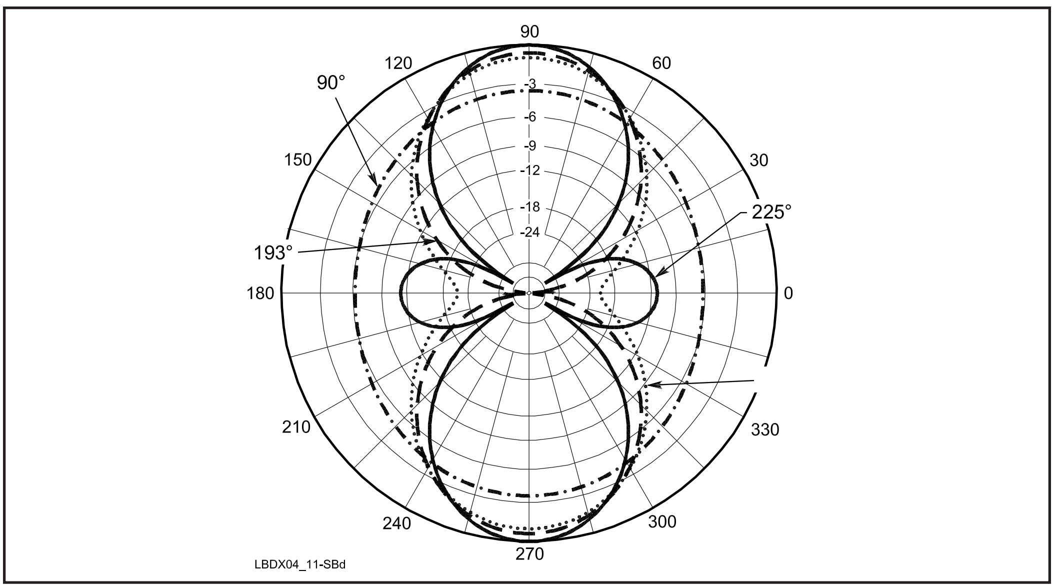



**Fig D—Azimuthal pattern (at 20° elevation angle) for broadside operation with variable spacing between the two elements. Note the sizeable sidelobe that appears for the 225° (5λ/8) case.**

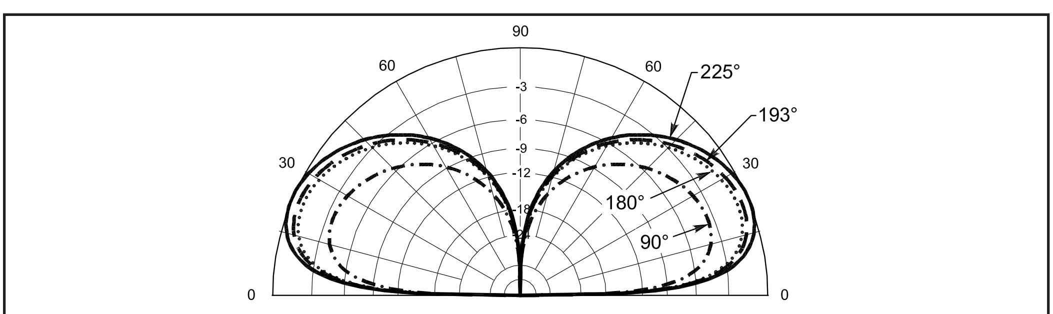



**Fig E—Elevation-plane patterns for different physical spacings between the 2-elements in a broadside array.**

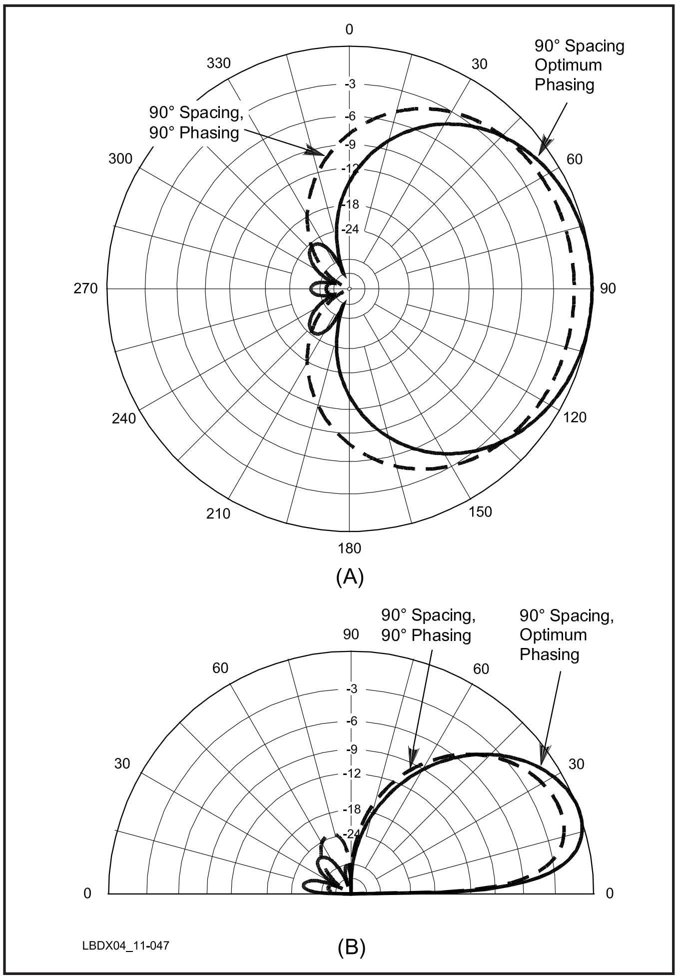



Fig 11-47—At A, solid line shows azimuth pattern (at 20° elevation) for quadrature-fed, 3-element in-line end-fire array, with spacings of λ/4 (Fig 11-46). Dashed line is for array fed with optimized phase angles and amplitudes. At B, elevation pattern comparisons.

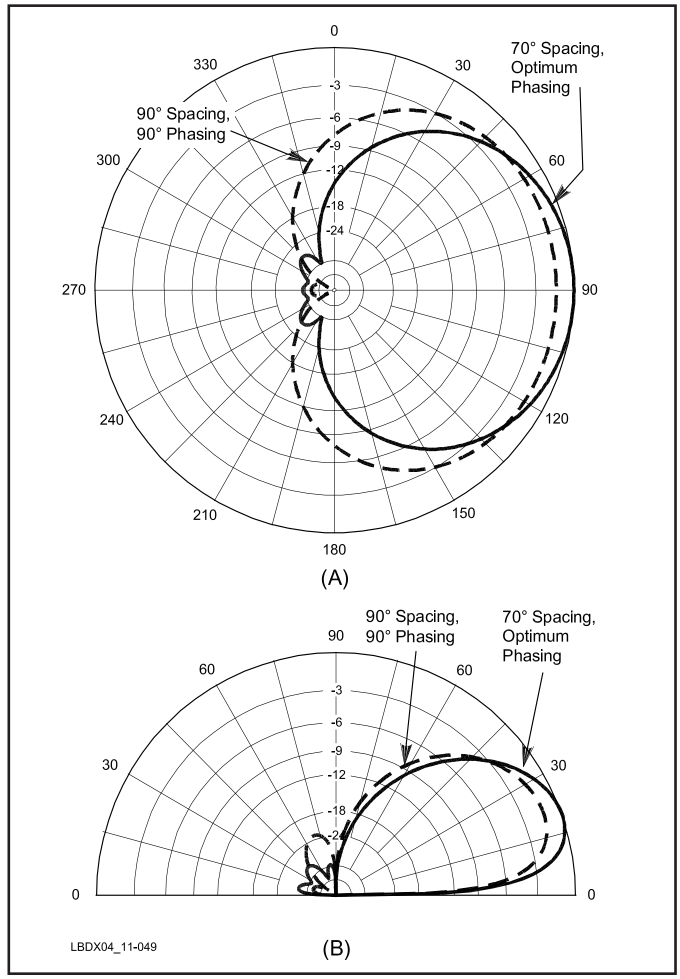



Fig 11-49—At A, solid line shows azimuth pattern (at 20° elevation) for Lahlum/Lewallen feed-optimized array using 70° spacings. Dashed line is reference with 90° spacings and 90° and 180° phasing. At B, elevation pattern comparisons.

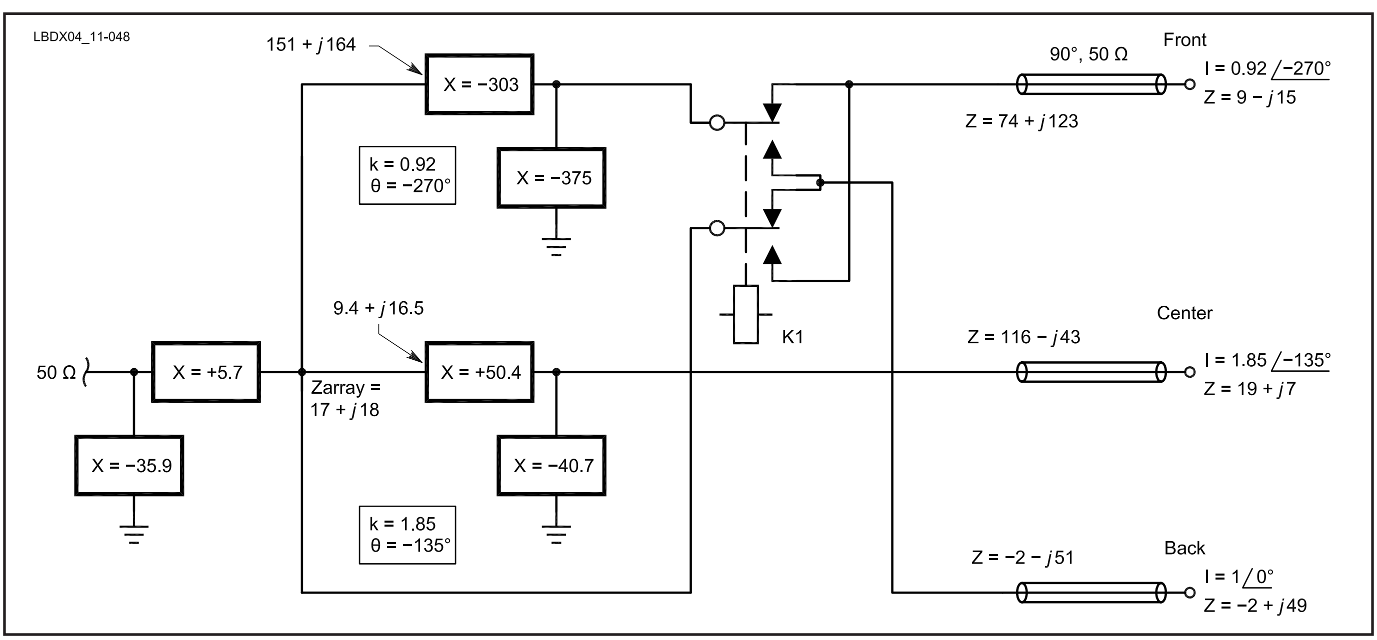



Fig 11-48—Lahlum/Lewallen feed network for a 3-element in-line, end-fire array with 70° spacing between the elements. This element phasing was chosen to be able to use λ/4 current forcing feed lines (Vf = 0.8). Direction switching is included.

**Table 11-12**
**Data, 3-element broadside arrays**

| *Spacing* | *Feed Currents* | *Gain dBi* | *Beamwidth 3-dB* | *RDF dB* | *DMF dB* | *Feed Impedances (Back, Mid, Front)* |
|---|---|---|---|---|---|---|
| 90° | 1, 0°; 2,–90°, 1, –180° | 5.28 | 143° | 9.24 | 17.9 | 15 – *j* 23; 26 – *j* 1; 77 + *j* 50 |
| 90° | 1, 0°; 1.75, –125°; 0.9, –250° | 6.59 | 106° | 10.9 | 27.9 | 11 – *j* 14; 26 + *j* 9; 30 + *j* 60 |
| 70° | 1, 0°; 1.85, –135°; 0.92, –270° | 6.57 | 98° | 11.2 | 27.8 | 9 – *j* 15; 19 + *j* 7; –2 + *j* 49 |
| 45° | 1, 0°; 1.9, –150°; 0.95, –300° | 5.19 | 91° | 11.5 | 27.5 | 5 – *j* 17; 11 + *j* 1; –18 + *j* 11 |

a certain attraction, since they make it possible to use the hybrid coupler (Collins) feed system.

### 4.4.1 Data, 3-element end-fire arrays

Note the negative impedance for the 70°-spacing case in **Table 11-12**. This happens frequently in multi-element arrays for the element in the back, especially at close spacings.

Note that with close spacing, especially at λ/4, the feed impedances become very low, which results in small bandwidth, critical tuning and less gain ($R_{rad}$ becomes small while $R_{loss}$ remains constant at 2 Ω).

### 4.4.2 Feed systems, 3-element end-fire arrays

### 4.4.2.1. Hybrid-coupler feed, 3-element end-fire arrays

The λ/4-spaced non-optimized version of this array can be fed with a hybrid coupler. **Fig 11-46** shows the feed system and direction switching and **Fig 11-47** shows the horizontal and vertical radiation patterns. As we need double the current in one of the elements of such an array, all we need to do is to run a coaxial cable with half the impedance of the coax feeding the other elements. In other words, the feed line to the center element will consist of two parallel-connected feed lines.

The transformed impedance for the center element (now being fed via a 270° long 25-Ω line) is 20.2 + *j* 8.4 Ω. The impedance at the end of the feed line going to the front elements is: 22.8 – *j* 18.4 Ω. To the back element: 49.8 + *j* 76.3 Ω (all calculated with the COAX TRANSFORMER/SMITH CHART software module). In parallel, those two give: 26.9 – *j* 10.1 Ω.

Notice that both impedances result in a low SWR in a 25-Ω system. The performance of the coupler will be very good if we design the hybrid coupler with a nominal impedance of 25 Ω. The values of the coupler components are:

$X_{L1} = X_{L2} = 25\ \Omega;\ X_{C1} = X_{C2} = 2 \times 25 = 50\ \Omega$

### *4.4.2.2. Lahlum-Lewallen feed, 3-element broadside array*

The array with 70° spacing between the elements has the advantage of not requiring 3λ/4 current-forcing feed lines if we use coaxial lines with a velocity factor of 0.8.

**Fig 11-48** shows the feed network, including the direction switching using a DPDT relay K1. **Fig 11-49** shows the radiation patterns for this feed-optimized array. Here, 50-Ω feed lines were used since they prevent the components in the L-network to the front element from having too high an impedance. A similar network can be calculated for other spacings and phase angles, using the *Lahlum.xls* spreadsheet and the appropriate NEW LOW BAND SOFTWARE modules. The procedure to adjust the L-network values is covered in Sections 3.6.1 and 3.6.2.

## 4.5. A Bidirectional End-Fire Array

Assume we have a 2-element broadside array with λ/2

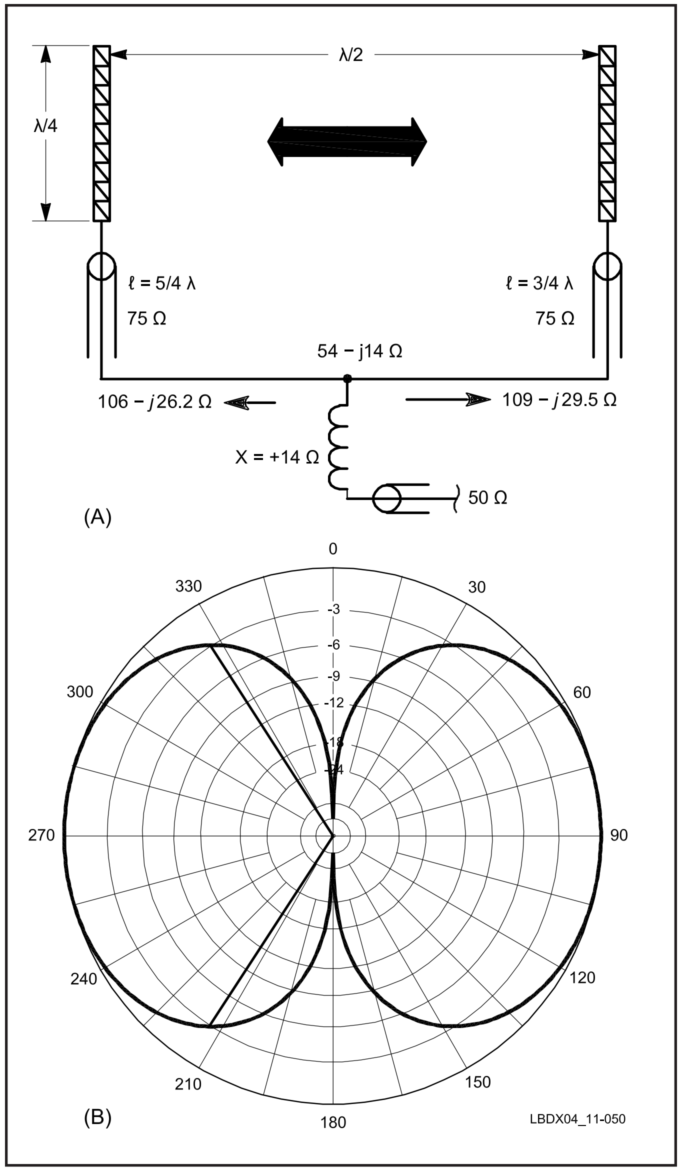



**Fig 11-50—Horizontal radiation pattern (at a 20° elevation) for the 2-element out-of-phase, end-fire array with λ/2 spacing. Elements in 90°-270° plane.**

spacing. How can we cover the 90° off directions? This can be done by feeding the two elements 180° out-of-phase, which also results in a bidirectional pattern but with a much broader lobe (beamwidth of 116° vs 64° in broadside) and less gain (3.5 dBi vs 4.9 dBi). See **Fig 11-50**.

### 4.5.1. Data, bidirectional end-fire array

Spacing: λ/2
Feed currents: I1 = 1 /0° A; I2 = 1 /– 180° A
Feed point impedance: Z1 = Z2 = 45 + *j* 14 Ω
Gain (over average ground): 3.51 dBi

### 4.5.2. Current-forcing feed system, bidirectional end-fire array

We will run a 270°-long (3λ/4) feed line to the element with the leading current, and a 450°-long (5λ/4) feed line to the element with the lagging feed current. With the lines being odd multiples of λ/4 long, we can use the current-forcing principle. A 90° and a 270°-long feed line are physically too short for the array, since the elements are spaced λ/2. To preserve symmetry, the T junction where the lines to the elements join must be located at the center of the array.

The impedances at the end of the feed lines can be calculated with the COAX TRANSFORMER software module. Using 75-Ω coax and zero losses we have: Z1′ = Z2′ = 114 – *j* 35 Ω. The combined impedance is 57 – *j* 17.5 Ω.

If we do the calculation including cable losses (there is a lot of cable in the two feed lines), assuming 0.2 dB/100 feet at 1.8 MHz and Vf = 0.66, we would have a feed impedance of 54 – *j* 14 Ω, which is a good match. In both cases we can tune out the negative reactance with a small series coil, and end up with a feed impedance very close to 50 Ω. See Fig 11-51.

## 4.6. Triangular Arrays

The original description by D. Atchley, W1CF, was for a 3-element array, where the verticals were positioned in an equilateral triangle with sides measuring 0.29 λ, or 104°. (Ref 939 and 941). The original version of the array used equal current magnitude in all elements. Later, Gehrke, K2BT, improved the array by feeding the two back elements with half the current of the front element. This very significantly improved the directivity of the array.

We can operate a triangle array in two different configurations:

- Beaming off the top of the triangle. The top corner (the front element) is fed with a phase delay vs the two bottom-line verticals, which are fed with the reference phase angle (0°)
- Beaming off the bottom of the triangle. In this case the bottom-corner elements are fed by the current with a phase delay vs the top vertical (the back element), which is fed with the reference phase angle of 0°.

In both cases the solitary element is usually fed with twice (or slightly less) the current magnitude when compared to the two non-solitary elements of the triangle, which are fed with the same current magnitude. Being a triangle,

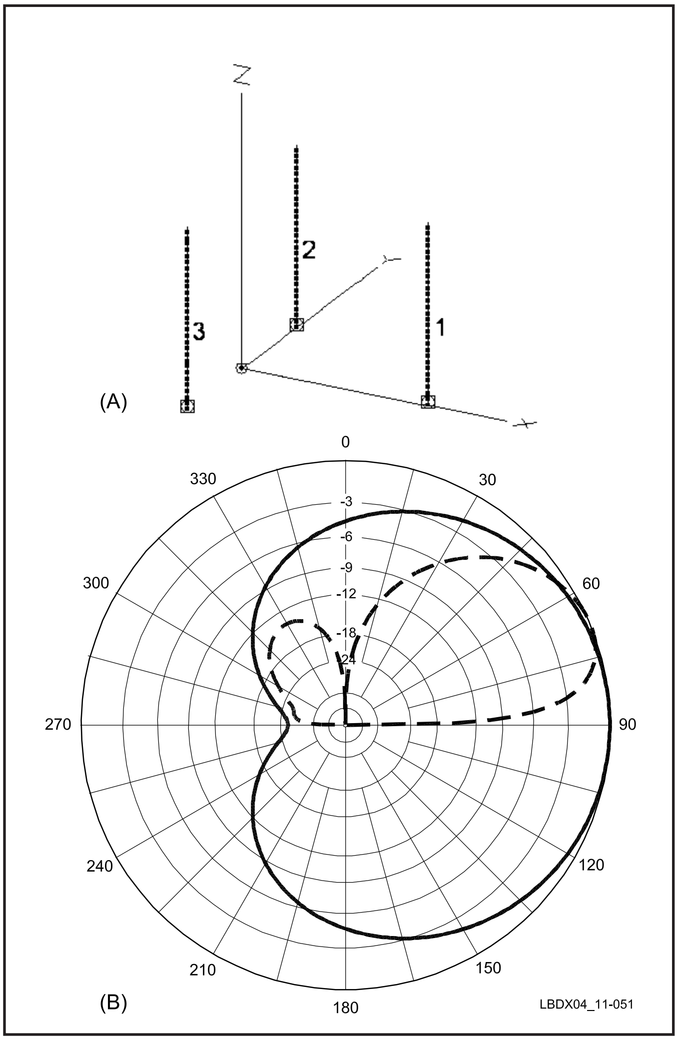



**Fig 11-51—Triangular array with 0.29-λ spacings between elements. Azimuth plot is at 20° elevation angle. See Table 11-13.**

**Table 11-13**
**Triangular Array Data**

| *Side* | *Config* | *Feed Current* | *Gain dBi* | *BW* | *RDF dB* | *DMF dB* | *Feed Impedance Front, Back* |
|---|---|---|---|---|---|---|---|
| 0.29λ | A | 2, –90°; 1, 0°; 1, 0° | 4.87 | 150° | 8.68 | 13.7 | 55 + *j* 19; 13 – *j* 36 (2x) |
| 0.29λ | A | 1.8, –110°; 1, 0°; 1, 0° | 5.47 | 129° | 9.40 | 16.0 | 53 + *j* 17; 13 – *j* 21 (2x) |
| 0.29λ | B | 2, 0°; 1, –90°; 1, –90° | 5.01 | 146° | 8.82 | 14.3 | 87 + *j* 0 (2×); 18 – *j* 9 |
| 0.29λ | B | 1.8, 0°; 1, –110°; 1, –110° | 5.56 | 126° | 9.49 | 16.3 | 76 + *j* 9 (2×); 14 – *j* 2 |

Side = side dimension in degrees (90° = λ/4)
Config A = shooting of the top of the triangle, B = shooting off the base

each array can be switched in three directions. Three directions fire off the top of a triangle, the other three off the bottom-line of a triangle. This means that a triangular array can be made switchable in six directions. All directions have the same gain (within 0.1 dB) and a very similar radiation pattern.

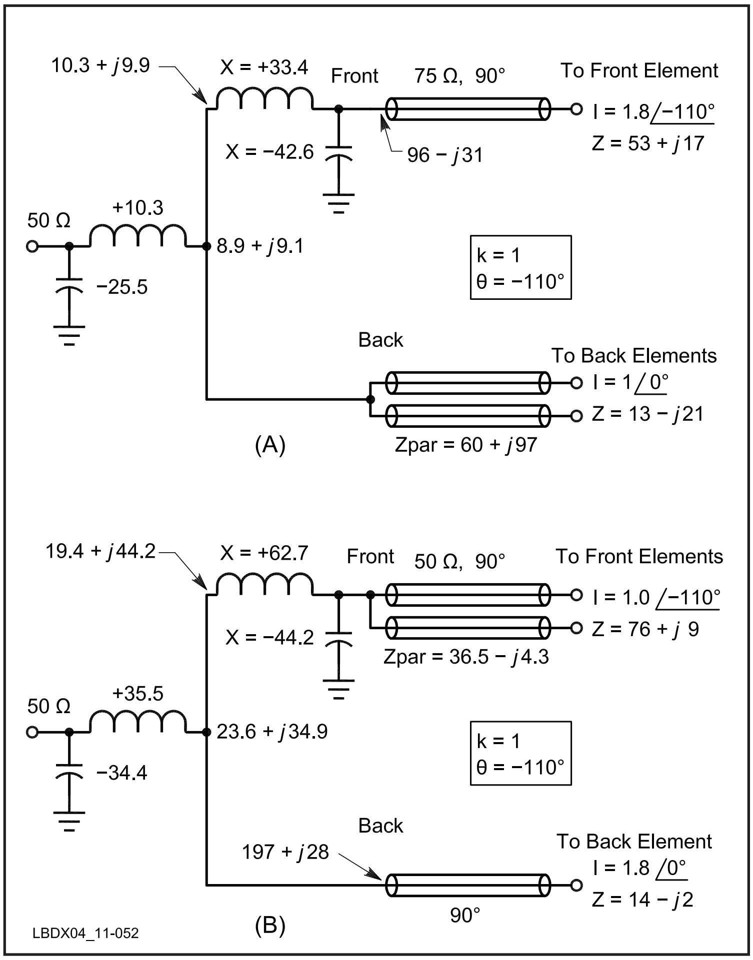



**Fig 11-52—At A, the feed system for the triangle array when firing off the top of the triangle. At B, the feed system when firing off the base line of the triangle. If you want six directions, you will need a switching system that selects the proper network, as shown in Fig 11-53.**

As expected, the performance (gain, beamwidth, directivity) is somewhere between the 2-element end-fire array and the Four-Square array. See **Table 11-13** and Fig 11-5**1.**.

### 4.6.1. Feed systems, triangular arrays

If we use quadrature feeding through a hybrid coupler, we are confronted with a practical switching problem. We need to feed the solitary element with double the feed current, which means with two paralleled feed lines. This means that a bunch of relays will be required to switch the extra feed line in parallel, depending on the direction.

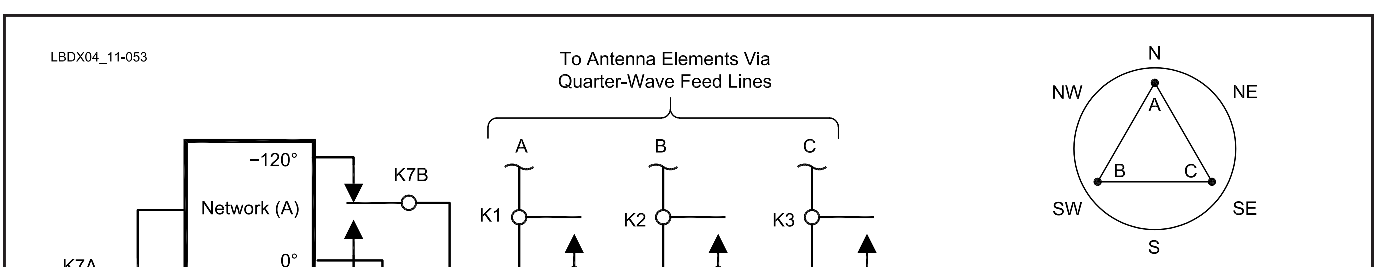
If you want to erect a triangle array, you should opt for the current-optimized Lahlum/Lewallen version, where you can achieve the double feed current magnitude by simply dimensioning the L-network components correctly. **Fig 11-52** shows the Lahlum/Lewallen feed networks for both triangle configurations.

**Fig 11-53** shows the direction switching for the array. As the feed impedances are different for the "A" and the "B" directions, we need two phasing networks. To do the direction switching we need a small matrix of SPST relays plus a seventh relay with three inverting contacts. This may seem complicated but using the two L-networks makes it possible to adjust the values to obtain the exact feed currents required. The measuring set up as described in Sections 3.6.1 and 3.6.2 should be used to make the adjustments.

## 4.7. The Four-Square Array

In 1965 D. Atchley, (then W1HKK, later W1CF, now a Silent Key), described two arrays that were computer modeled, and later built and tested with good success (Refs 930, 941). Although the theoretical benefits of the Four-Square were well understood, it took a while before the correct feed methods were developed that could guarantee performance on a par with the theory.

The Four-Square is in fact similar to a 3-in-line




| XX | K1 | K2 | K3 | K4 | K5 | K6 | K7 |
|---|---|---|---|---|---|---|---|
| N | X | O | O | I | X | X | O |
| NE | X | O | X | I | X | O | X |
| SE | O | O | X | X | O | X | O |
| SE | O | X | X | X | O | O | X |
| SW | O | X | O | X | O | O | O |
| NW | X | X | O | O | X | X | X |

**Fig 11-53—Seven relays, of which six are SPST relays in a matrix, are used to make a 6-direction switching network/feed system. The networks are shown in Fig 11-52.**

end-fire array—the center two elements are fed in-phase and act as one common element. If all four elements have equal current, the total center-element current (for both in-phase elements together) is twice the current at each end. The required 1:2:1 current distribution as explained in Section 4.4 is satisfied.

The Four-Square can be switched in four quadrants. Atchley also developed a switching arrangement that made it possible to switch the array directivity in increments of 45°. The second configuration consists of two side-by-side cardioid arrays. This antenna is discussed in detail in Section 4.8.

The practical advantage of the extra directivity steps, however, does not seem to be worth the effort required to design the much more complicated feeding and switching system, since the forward lobe is so broad that switching in 45° steps makes very little difference. It is also important to keep in mind that the more complicated a system is, the more failure-prone it is.

### 4.7.1. Quadrature-fed, λ/4-spaced Four-Square

Placement of elements is in a square, spaced λ/4 per side. All elements are fed with equal currents. The back element is fed with the reference feed current angle of 0°, the two center elements with −90° phase, and the front element with −180° phase difference.

**Fig 11-54** shows the radiation patterns for this array. The direction of maximum signal is along the diagonal from the rear to the front element. An array always radiates in the direction of the element with the lagging current.

#### *4.7.1.1. Data, λ/4-spaced Four-Square array*

Dimension of square side: λ/4.
Feed currents:

I1 = 1 /–180° (front element)
I2 = I4 = 1 /–90° (center elements)
I3 = 1 /0° (back element)
Gain: 6.67 dBi over good ground
3-dB beamwidth: 98°
RDF = 10.58 dB
DMF = 21.02 dB

Feed-point impedances:

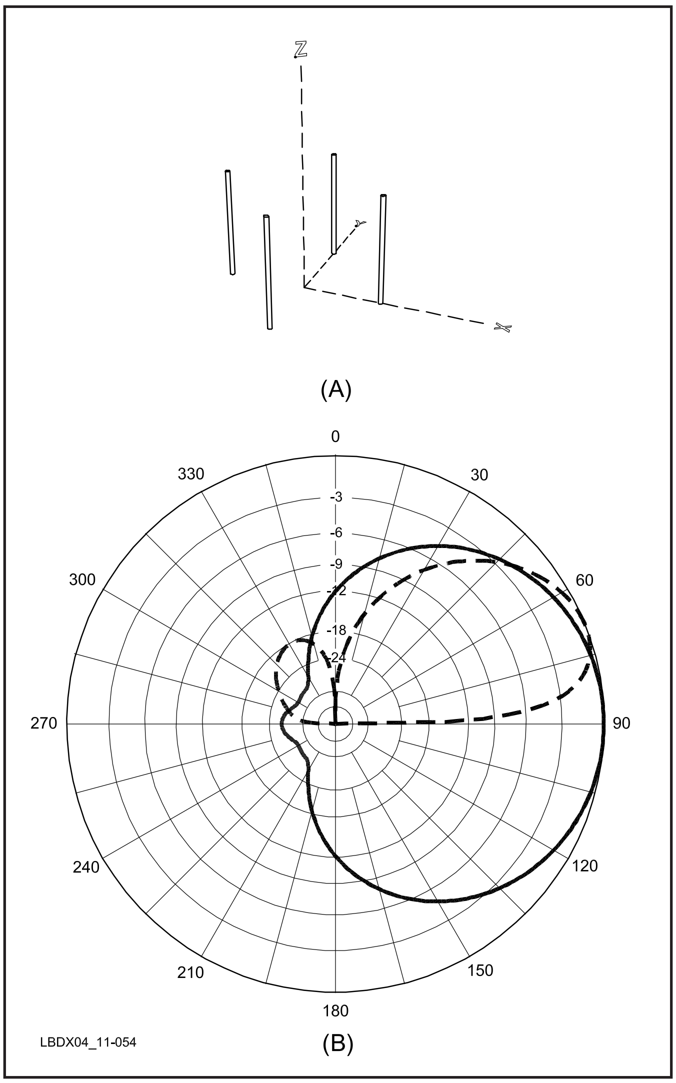



**Fig 11-54—Radiation patterns (horizontal at a 20° elevation angle) for a typical quadrature-fed Four-Square array. Notice the important back lobe at relatively high elevation angles (about 60°).**

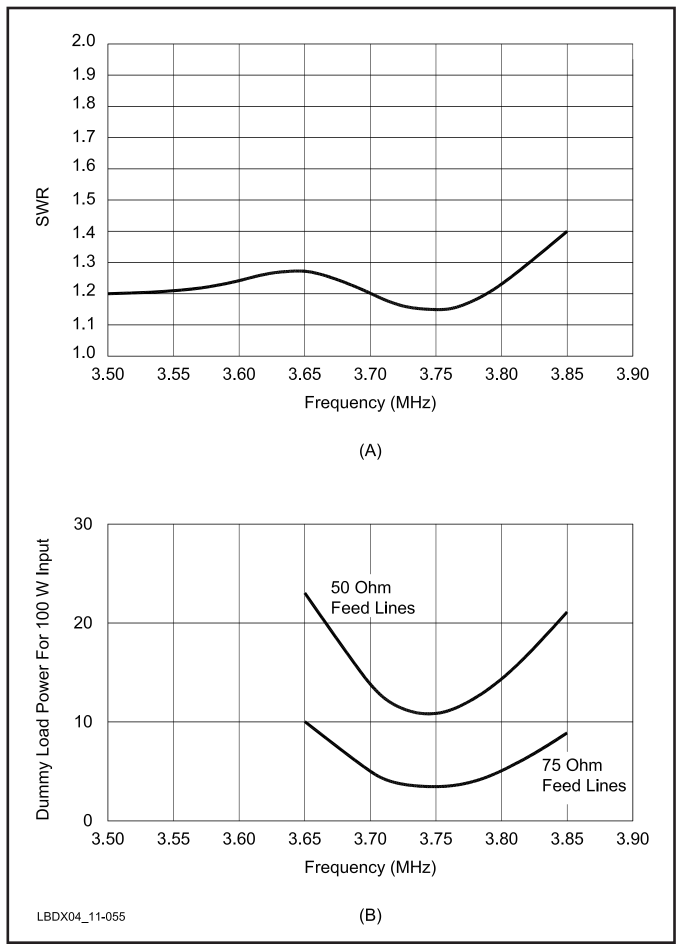



**Fig 11-55—SWR and dissipated-power curves for a Four-Square array tuned for operation in the 3.7 to 3.8-MHz portion of the 80-meter band. Note that the dissipated power is much lower with 75-Ω feed line than with the 50-Ω feed line. The SWR curves for both the 50- and the 75-Ω systems are identical. The curve remains very flat anywhere in the band, but it is clear that the power dissipated in the load resistor is what determines a meaningful bandwidth criterion for this antenna.**

Z1 = 62.3 + *j* 53.4 Ω
Z2 = Z4 = 40.5 – *j* 19 Ω
Z3 = –0.3 – *j* 15.2 Ω

### *4.7.1.2. Feed system, quadrature–fed Four-Square*

#### *4.7.1.2.1. Hybrid-coupler Collins feed, quadrature-fed Four-Square*

Since the antenna is fed in quadrature, a hybrid feed system is possible (see Section 3.4). We can feed the array with either 50 or 75-Ω, λ/4 current-forcing feed lines. Using 75-Ω feed lines generally results in less power being dumped in the load resistor if the hybrid network is designed for a system impedance of 50 Ω. On the antenna design frequency it should be possible to dump no more than 1% to 5 % (–20 to –13 dB) of the transmit power in the dummy load. A 200-W dummy load should normally be sufficient for 1.5 kW power output into the antenna. See **Fig 11-55**. It's not a bad idea, however, to have a bigger one. In case of malfunction of the antenna much more power can be dumped into the load! Many operators measure the power dumped in the dummy and have an indicator in the shack.

#### *4.7.1.2.2. Lewallen feed, quadrature-fed Four-Square*

In Section 3.4.5 we see the detailed calculation of the Lewallen feed system (LC-network) using the *Lahlum.xls* spreadsheet. The Lewallen feed method for this array is worked out in great detail in *The ARRL Antenna Book*, where L-network values are listed for a range of feed-line impedances and ground systems.

### 4.7.2. WA3FET-optimized Four-Square array

Jim Breakall, WA3FET, optimized the quarter-wave-spaced Four-Square array to obtain higher gain and better directivity. Fig 11-54 shows that the original Four-Square exhibits a big high-angle backlobe (down only 15 dB at 120° in elevation). By changing the feed current magnitude and angle to the various elements you can change the size and the shape of the backlobes as well as the width of the front lobe. Full optimization is a compromise between optimization in the elevation and the azimuth planes. With Breakall's optimization, the gain of the array goes up by 0.6 dB. At least as important is a significant gain in directivity (RDF and DMF).

#### *4.7.2.1 Data, WA3FET-optimized Four-Square*

Dimension of square side: λ/4
Feed currents: I1 = 0.872 /–218° A (front)
I2 = I4 = 0.9 /–111° A
I3 = 1 /0° A (back)
Gain: 7.25 dBi
3-dB beamwidth: 85°
RDF = 11.4 dB
DMF = 24.4 dB
Feed-point impedances:
Z1 = 37.5 + *j* 57.7 Ω (front)
Z2 = Z3 = 30.8 – *j* 7.0 Ω (center)
Z4 = 6.0 – *j* 3.4 Ω (back)

### 4.7.3. Lahlum/Lewallen feed system, quadrature-fed Four-Square

In Section 3.4 I covered in detail the design of the Lahlum/Lewallen feed system for this array. Note that we lengthened all elements an equal amount to obtain a non-reactive impedance in the center two elements, resulting in slightly different component values and impedances.

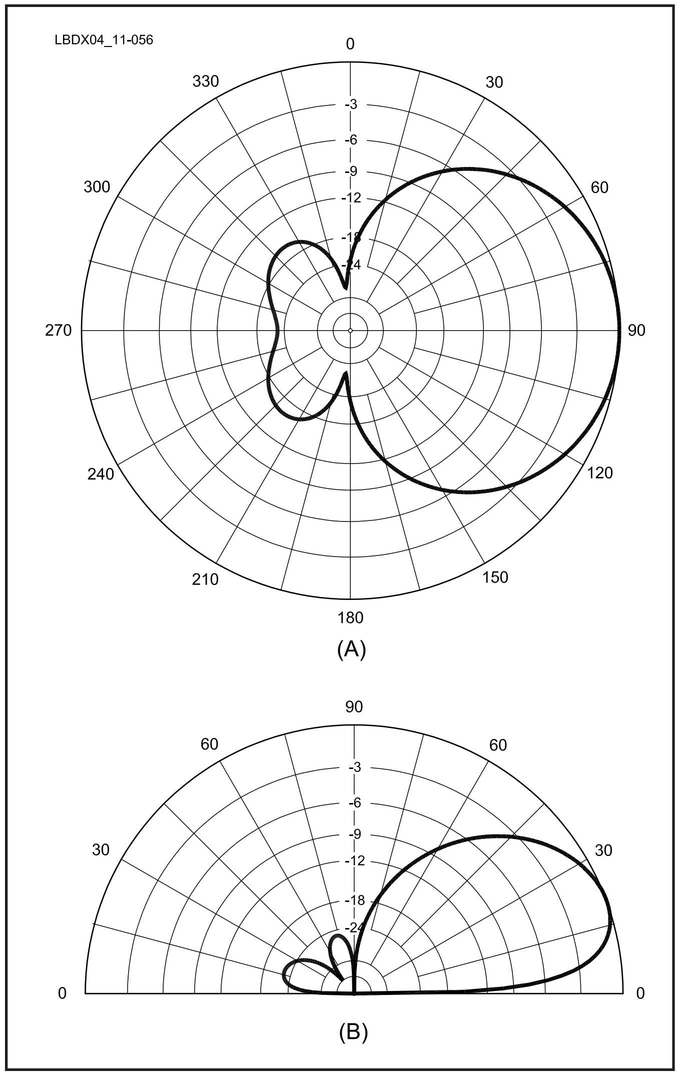



**Fig 11-56—Radiation patterns for the WA3FET-optimized Four-Square, where the high-angle back lobe has been reduced substantially. Net result is 0.7 dB more gain and increased directivity.**

### 4.7.4. W8JI cross-fire feed, Four-Square

W8JI's 4-square has the following configuration:

Dimension of square side: λ/4
Feed currents: I1 = 1 /–240° (front element)
I2 = I4 = 1 /–120° (center elements)
I3 = 1 /0° (back element)

If you model this configuration you find:

Gain: 7.45 dBi (0.8 dB better than quadrature-fed)
3-dB beamwidth: 79°
RDF = 11.78 dB
DMF = 17.4 dB
Feed-point impedances:
Z1 = 27 + *j* 56 Ω
Z2 = Z4 = 24 Ω
Z3 = 6.6 + *j* 3 Ω

The impedance of 24 Ω (at 1.83 MHz) was obtained by

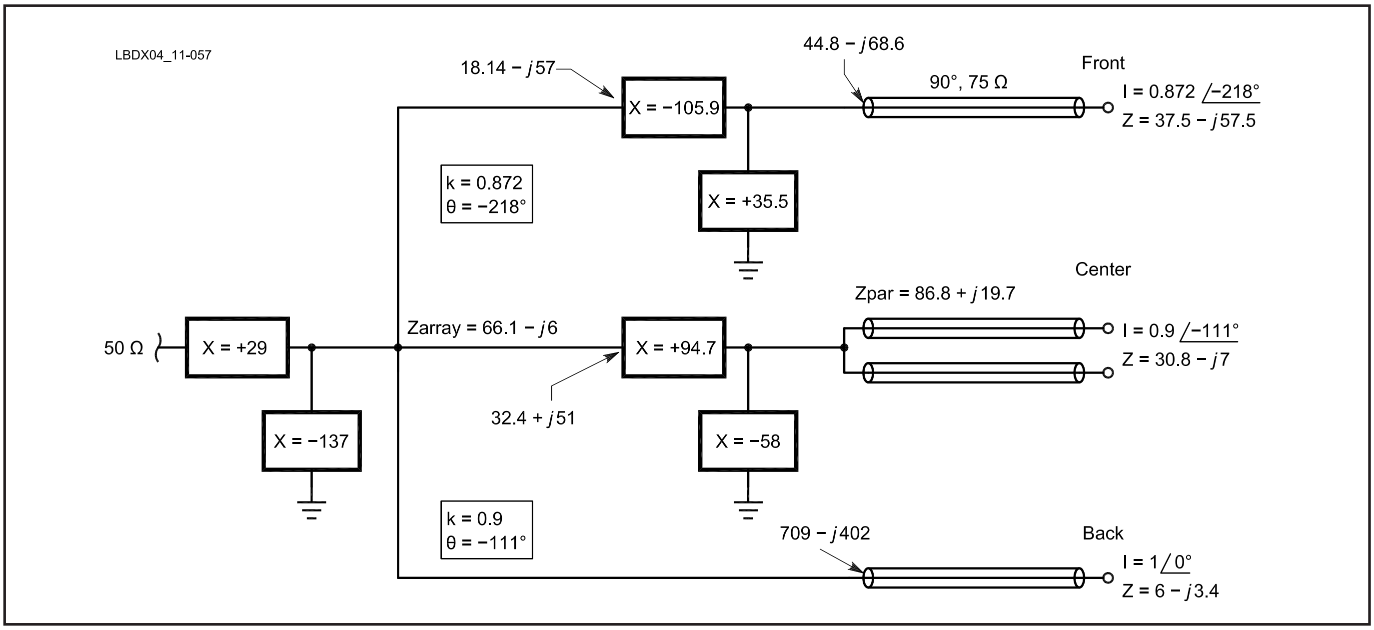



**Fig 11-57—Lahlum/Lewallen feed circuit for the WA3FET-style Four-Square, with optimized phase angle and drive current magnitudes. In Fig 11-19 slightly different element impedances were used. Note that the variation of the L-network components are well within the normal tuning range. The feed impedances are also within a few percent of one another.**

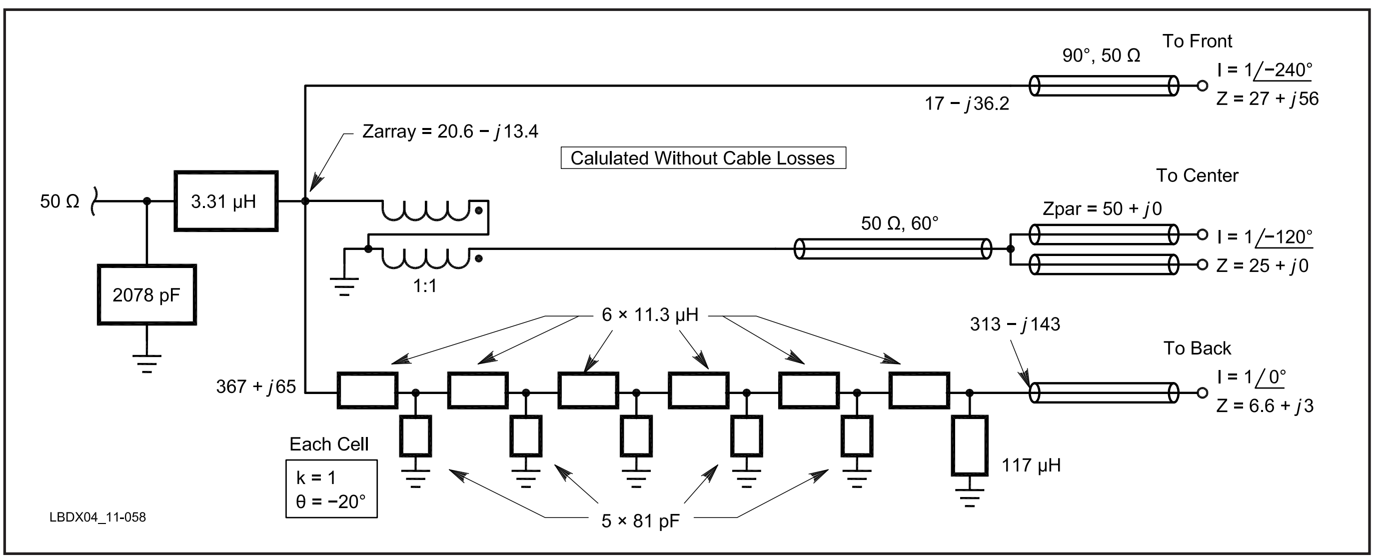



**Fig 11-58—Feed system used by W8JI for his 160-meter Four-Square, which uses 120° phasing angle increments. See text for details.**

tuning the solitary elements for resonance at 1.818 MHz.

**Fig 11-58** shows the feed system developed by Tom, W8JI. The cross-fire principle means that we feed the array from the front element, and use a 180° phase-reversal transformer to feed the other elements (see Section 3.4 and also Chapter 7).

With respect to the front element, the required phase shift to the center elements is +120°, which is equal to +120–360 = –240°. Note that the parallel impedance of the λ/4 current-forcing feed lines to the two center elements is very close to 50 Ω. This means that we will be able to obtain any desired phase shift by using a 50-Ω feed line of a length (in degrees) equal to the required phase shift.

Note the 180° phase-reversal transformer, which takes care of –180° of the required 240° phase shift. The remaining 60° is obtained through a 50-Ω coax 60° in length. The feed system for the front and the center elements is extremely broad-banded, as phase shift remains constant with changing frequency, because of the cross-fire principle.

The "bad boy" is the back element. We can feed it with an L-network and use the *Lahlum.xls* spreadsheet to calculate the components. It is obvious that this branch will be the bottleneck for bandwidth. You could develop two L-networks, one for each band section of interest, and switch them, however.

Tom, W8JI, used what he calls "an artificial transmission line using lumped components, composed of multiple L/C sections to simulate a transmission line with a characteristic impedance matching the rear element." Tom quotes the following advantages: "*Q is low, making phase-shift much less frequency critical. and I can easily tweak delay-line characteristics with a few adjustments to optimize the array null*."

How do you calculate such an artificial line? You can consider it as a series connection of a number of L-networks—which we know from the Lewallen/Lahlum principle. Here is how to calculate the components of the "artificial transmission line" using *Lahlum.xls*:

- First calculate the impedance at the end of the λ/4 feed line to the back element: 313 – *j* 143 Ω.
- Next use this impedance as an input for R and X in the second part of the spreadsheet (called "for non-current forcing").
- For the regular Lahlum network composed of a single L-network cell, we would enter a required phase shift of (+240 – 360) = –120°, and end up with a parallel cap of 293 pF and a series coil of 28.5 µH (all calculated for 1.83 MHz). The input impedance into the L-network would be 94.6 + *j* 163.8 Ω.
- But you can specify, for example, a required phase shift of –20°. In that case this L-network cell will require a parallel coil of 117 µH and a series coil of 11.3 µH. Look now at the input impedance and note it is 367 + *j* 65 Ω.
- All we need to do now is, using the same spreadsheet, calculate another five L-networks, around an output impedance of 367 + *j* 65 Ω, each for a 20° phase shift. Each of these L-network cells has a parallel capacitor of 81 pF and a series coil of 11.3 µH, and the input impedance of this artificial line (consisting of six L-network cells), is 367 + *j* 65 Ω.

W8JI has experimented a lot with this system, and notes: "*Because the current is low, components can be modest sized. The end result is more bandwidth, more stability and less loss than a simple one-stage network.*" It is not strictly necessary to use six cells, of course, but the greater the number, the better the bandwidth.

Another way to calculate the "artificial transmission line" is to first tune out the reactance of the impedance at the end of the λ/4 feed line. A parallel coil of 77 µH, which represents +828 Ω reactance, will turn the impedance into 378 Ω. Now we can use the PI-LINE STRETCHER module from the NEW LOW BAND SOFTWARE to calculate cells that each give 20° phase shift and for a characteristic impedance of 378 Ω. The values of the components are identical.

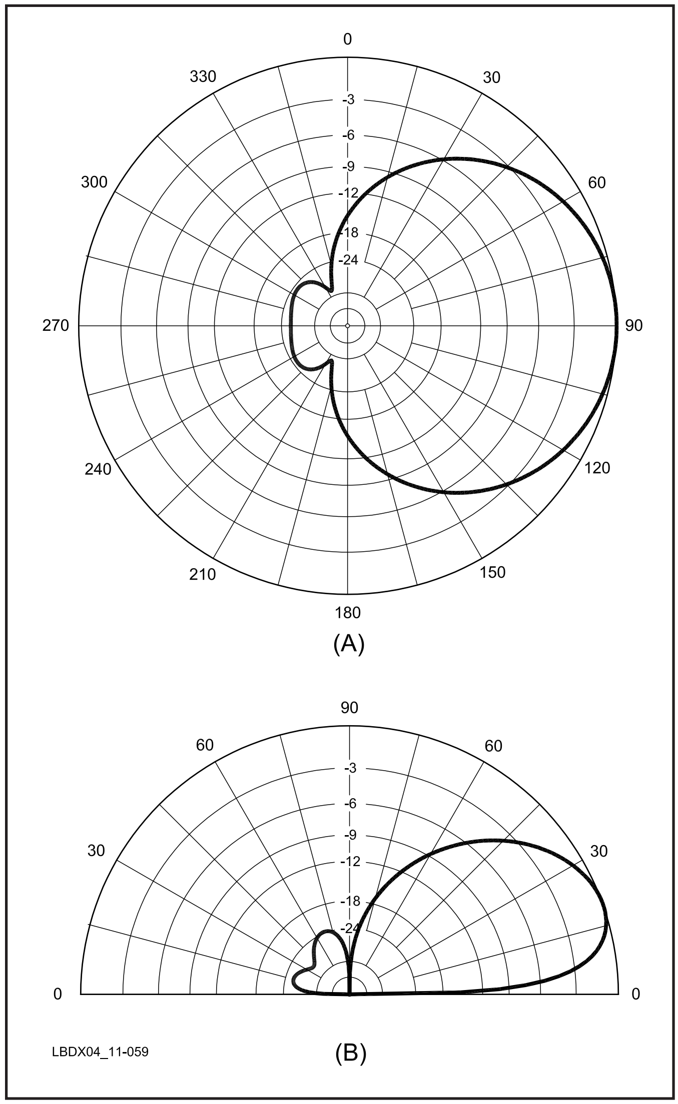



**Fig 11-59—Radiation patterns for a reduced-size (λ/8 side) Four-Square, which exhibits even better directivity than the larger varieties but with slightly less gain and less bandwidth.**

### *4.7.4.1. Other applications, cross-fire principle*

Single L-networks can be replaced with multiple-section networks to improve bandwidth. This is especially true where high-impedances are encountered, which is most frequently the case with the "back element" of an array.

#### *4.7.4.1.1. Where can we apply this cross-fire principle?*

To be able to feed an element through a 1:1 (180°) transformer and a coaxial phasing line (whose length is 180° minus the required phase delay), you need to be able to achieve a pure resistive impedance at the end of the current-forcing feed line. You can shorten/lengthen the element somewhat so that the feed-point impedance at the element is purely resistive.

Next, you can select the impedance of the current forcing feed lines (usually 75 or 50 Ω), and see if any combination turns out to be a good one. Good ones are: 25 Ω (two 50 Ω in parallel), 30 Ω (50 and 75 Ω in parallel) 37.5 Ω (two 75 Ω in parallel), 50 Ω and 75 Ω.

It might be better to make an element somewhat non-resonant, so that with the addition of a parallel reactance (coil or capacitor) we end up near one of the above-mentioned impedances.

### *4.7.4.2. Conclusion, cross-fire principle*

It has to be rather a lucky shot if you can apply this principle. It is clear that the cross-fire principle may track frequency a little better than the other feed methods, and this is certainly so with receiving arrays (and phased Beverages) where the element impedances hardly change with frequency.

If you need bandwidth, it seems to me that the easiest solution is to provide switchable L-networks; that is, one for the CW end and one for the Phone end of the band, while at the same time changing the length of the current-forcing feed lines. (You would add some extra length for the lower frequency) and retune the elements for resonance.) That is certainly a guarantee for peak performance and at the same time it gives you the ability to prune the array for optimum performance, even at the sacrifice of some bandwidth.

### 4.7.5. The λ/8-spaced Four-Square

#### *4.7.3.1. Data, λ/8-spaced Four-Square*

Dimension of square side: λ/8
Feed currents:
I1 = 1 /–270° A
I2 = I3 = 1 /–135° A
I4 = 1.1 /0° A
Gain: 5.85 dBi
3-dB beamwidth: 89°
RDF = 11.3 dB
DMF = 25.0 dB
Feed-point impedances:
Z1 = – 11.3 + *j* 18.7 Ω
Z2 = Z3 = 18.4 – *j* 5.6 Ω
Z4 = 1.3 – *j* 11.8 Ω

This small-footprint Four-Square sacrifices 1.4 dB of gain compared to its optimized big brother, but it has every bit as good or even better directivity. The main disadvantage of this design is the much narrower bandwidth. Note that the reduction in gain is to a large extent due to the lower impedances of the elements, taking into account that we inserted an equivalent-ground-radials loss resistance of 2 Ω at the base of each element.

#### *4.7.5.1. Feeding the "small" Four-Square*

Because of the very low impedances involved, feeding this array is tricky and at best the bandwidth will be narrow. Let's take a close look at **Fig 11-60**. As usual we will feed the back element directly. Note that the negative impedance we've become accustomed to is large, which indicates very heavy mutual coupling, obviously due to the proximity of the elements involved. The center elements have reasonable impedance values, which translates into normal L-network components in the center branch. The branch to the front element is very peculiar. The real part of the impedance of the front element (including 2-Ω ground losses) is 1.3 Ω. This means that this element is taking almost no power at all. If we do the calculating of the Lahlum-network we will some up with an "extreme" value for the series element in the network (–4755 Ω), which represents 18 pF at 1.8 MHz. It is clear that due to stray capacity this is a impossible value. The value X1 (for the parallel element of the L-network) is the Lahlum-network value. Let's see what would be the value of a parallel impedance that turns 51.9 + *j* 471 Ω into a pure resistance. Using the HUNT-SERIES IMPEDANCE NETWORK module of the NEW LOW BAND SOFTWARE, it appears that it is –477 Ω, and that the resistive impedance at that point is 4,326 Ω, a high value as expected.

At this point is appears to be much simpler to turn the front element into a parasitic element and not feed it at all. The parasitic element can now be tuned by simply tuning the parallel reactance (a capacitor), which in this case has the value X2 = –477 Ω. Note also that X2 is almost the same as X1. If the front element was taking no power at all, these two value would have been identical.

In practice, you can just leave out the series element of the L-network in the front element branch. If you can, stay away from arrays with such close coupling and such low impedances. They mean critical alignment, high Q and low bandwidth!

### 4.7.6. Direction switching for Four-Square arrays

**Fig 11-61** shows a direction-switching system that can be used with all Four-Square arrays. The front element (in the direction of firing) will be fed with the most lagging feed angle (–180° for quadrature feeding); the back element will be with the zero reference feed angle. "Mid", Back" and

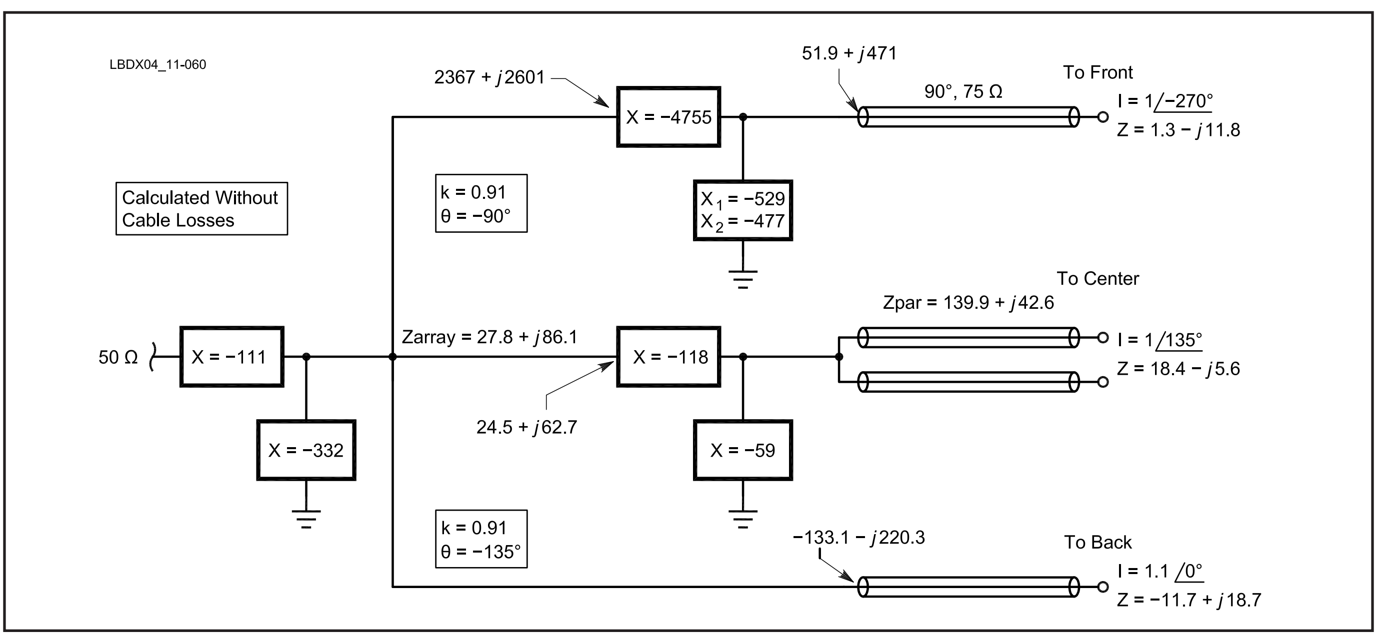



**Fig 11-60—Lahlum/Lewallen L-network feed system for the small 4-square with λ/8 side dimensions.**

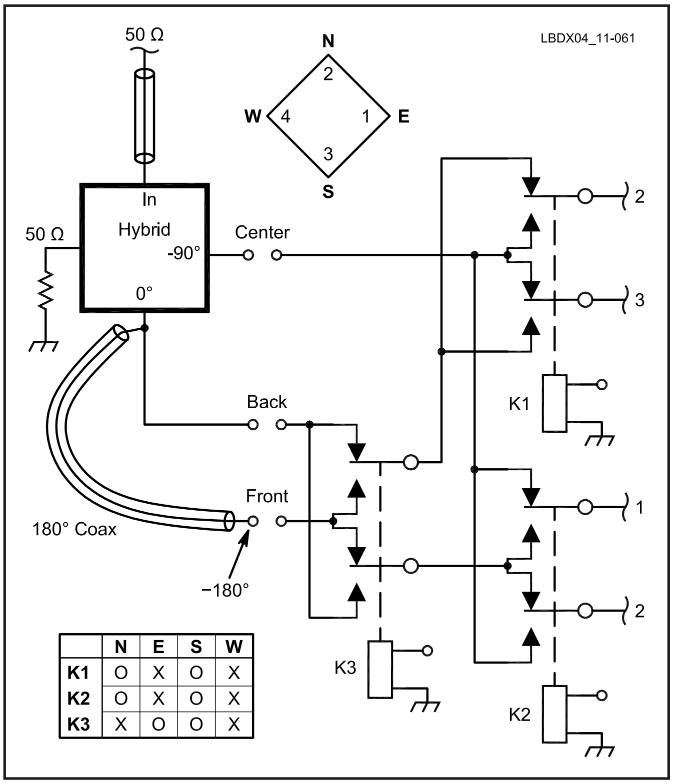



**Fig. 11-61—This direction-switching system shown can be used with λ/8 Four-Square arrays, with whatever phasing circuitry is used.**

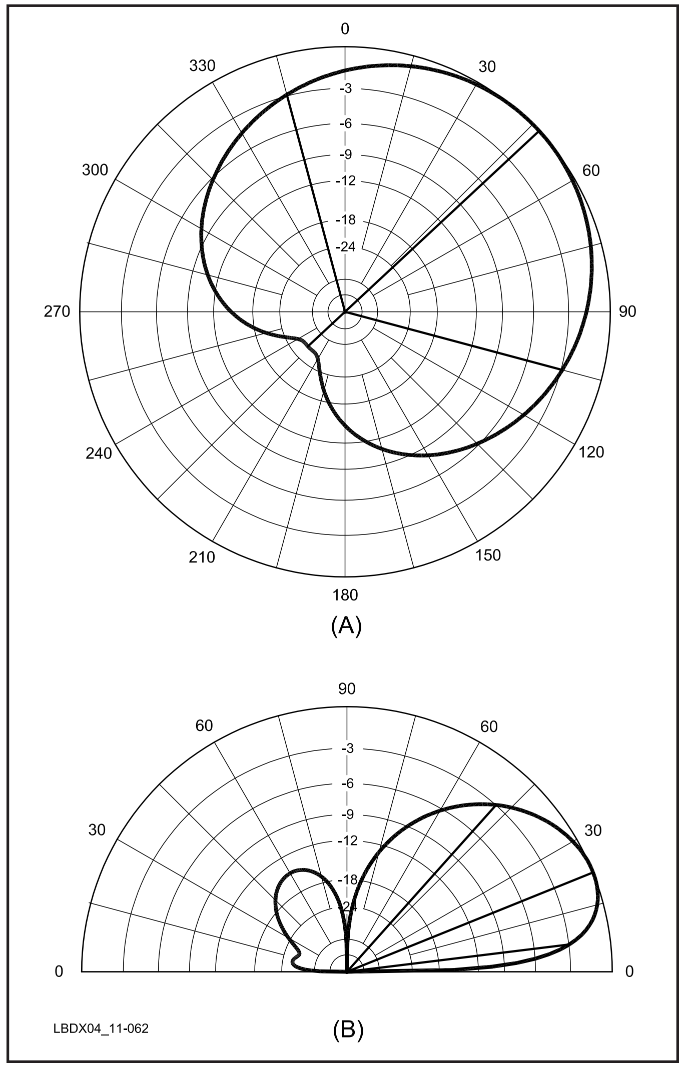



**Fig 11-62—Layout and radiation patterns (horizontal at 20° elevation angle) for the close-spaced side-by-side, end-fire cell configuration, used in an "intermediate" direction in a Four-Square. Compare with the patterns of a conventional Four-Square in Fig 11-54.**

"Front" go to the corresponding points in the feed circuits for a Lahlum/Lewallen circuit.

For a quadrature-fed array, a hybrid can be used, wired as shown in Fig 11-61. The Comtek Hybrid Coupler includes a 180° phase-inversion transformer. In that case "Front" is directly connected to that transformer output, without using the 180° phase-shift coax shown in the figure.

## 4.8. Four-Square Array with 8 Directions

A Four-Square usually has a −3-dB forward lobe beamwidth of 85° to 100° (depending on spacing and phasing), so four directions can quite adequately cover all azimuths. Some people think they need more. Admittedly, more than four directions may be advantageous for moving the nulls, which may be advantageous for nulling out QRM and noise from certain directions.

In the half-angle intermediate position, the four verticals are fed like two side-by-side (broadside is not an adequate term—"narrow-side" would be better) end-fire cells, spaced only λ/4. **Fig 11-62** shows the radiation patterns for a Four-Square using side-by-side, end-fire feeding.

### 4.8.1. Quadrature-fed, 8-direction Four-Square

#### *4.8.1.1. Data, quadrature-fed 8-direction Four-Square*

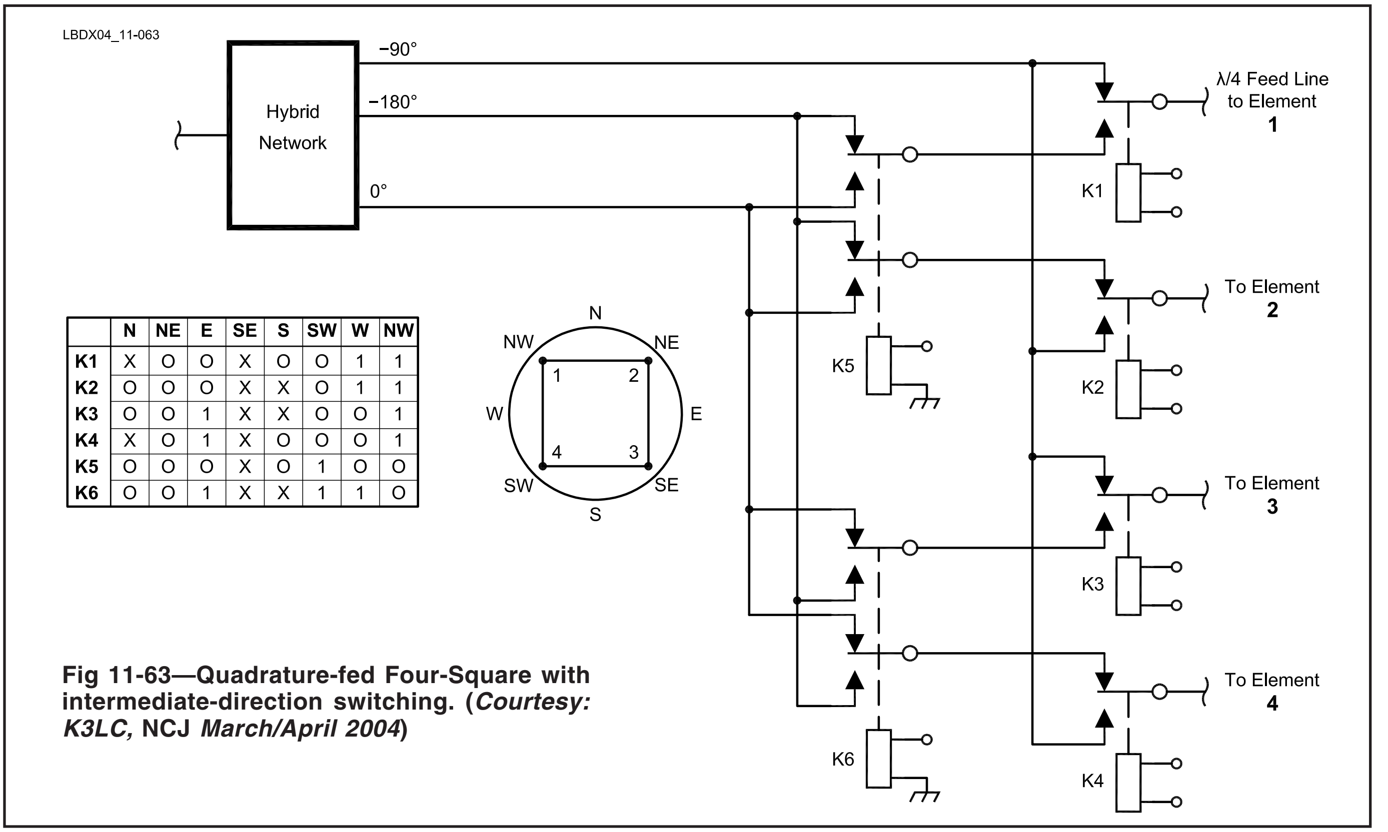
Side of square: λ/4
Feed currents:
I4 = I3 = 1 /–90° A (front)
I1 = I2 = 1 /0° A (back)
Gain: 5.51 dBi (6.67 dBi in regular Four-Square configuration)
3-dB beamwidth: 123°
RDF = 9.22 dB
DMF = 14.7 dB
Feed-point impedances:
Z4 = Z3 = 88.5 + *j* 6.6 Ω (front)
Z1 = Z2 = 17.7 – *j* 36.1 Ω (back)

#### *4.8.1.2. Feed system, quadrature-fed 8-direction Four-Square*

If you use quadrature feeding, you can feed the array in all eight directions using a hybrid coupler feed system. **Fig 11-63** shows a possible direction-switching and feed system for the quadrature-fed Four-Square with hybrid coupler, including the four additional "mid-direction" firing directions.

### 4.8.2. Optimized feeding, 8-direction Four-Square

For spacing of λ/4 between the elements, if we increase the phase shift to 105°, we get somewhat higher gain and directivity. For that case we need to feed the array with a Lewallen feed




| | N | NE | E | SE | S | SW | W | NW |
|---|---|---|---|---|---|---|---|---|
| **K1** | X | O | O | X | O | O | 1 | 1 |
| **K2** | O | O | O | X | X | O | 1 | 1 |
| **K3** | O | O | 1 | X | X | O | O | 1 |
| **K4** | X | O | 1 | X | O | O | O | 1 |
| **K5** | O | O | O | X | O | 1 | O | O |
| **K6** | O | O | 1 | X | X | 1 | 1 | O |

**Fig 11-63—Quadrature-fed Four-Square with intermediate-direction switching. (*Courtesy: K3LC,* NCJ *March/April 2004*)**

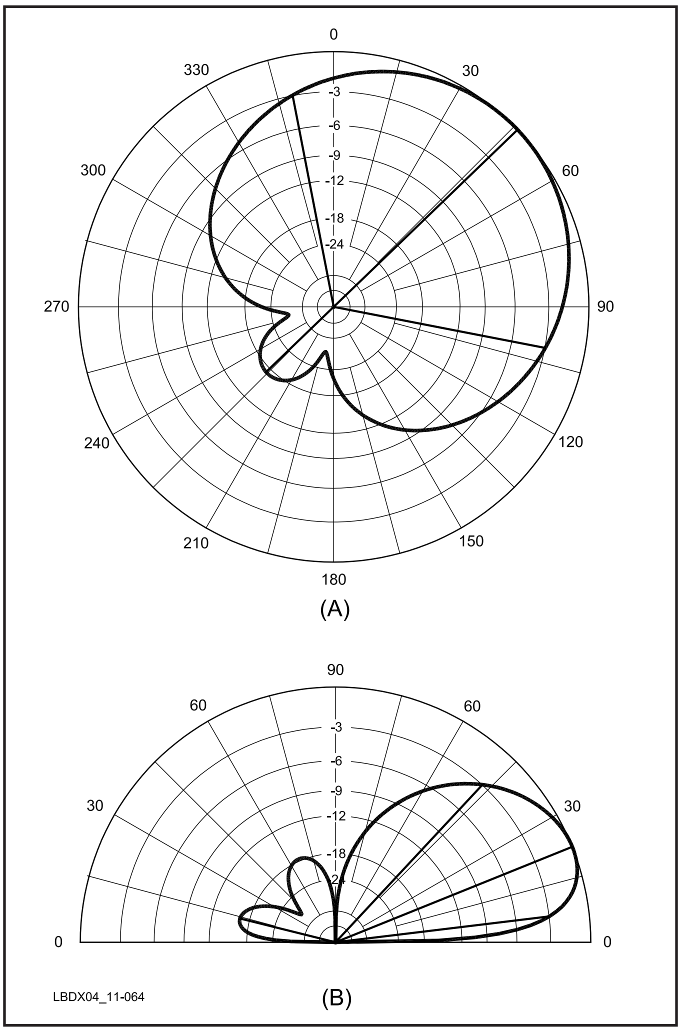



**Fig 11-64—Radiation patterns (horizontal at 20° elevation angle) for the optimized side-by-side, end-fire cells, showing improved high-angle rejection off the back due to 105° phase-shift feed at intermediate angles.**

system (L-network). If you decide to add four more directions to an optimized Four-Square, this is the way to go.

### *4.8.2.1. Data, optimized 8-direction Four-Square*

Side of square: λ/4
Feed currents: I4 = I3 = 1 /–105° A (front)
I1 = I2 = 1 /0° A (back)
Gain: 5.92 dBi (optimized regular Four-Square configuration: 7.25 dBi)
3-dB beamwidth: 113°
RDF = 9.72 dB
DMF = 16.5 dB
Feed-point impedances:
Z4 = Z3 = 81.8 + *j* 15 Ω (front)
Z1 = Z2 = 13.2 – *j* 26.3 Ω (back)

### *4.8.2.2. Feed system, optimized 8-direction Four-Square*

**Fig 11-65** shows the Lahlum/Lewallen feed system for the Four-Square in its intermediate directions. Note that we use 75-Ω feed lines, which results in a relatively high feed impedance for the array. In this case a simple coil in parallel with the input ( X = +113.9 Ω) will turn the feed impedance into 52 Ω, a perfect match!

If we choose to feed the back elements directly and the front elements through a 105° L-network delay circuit, we will have L-network components of lower impedance (Xp =

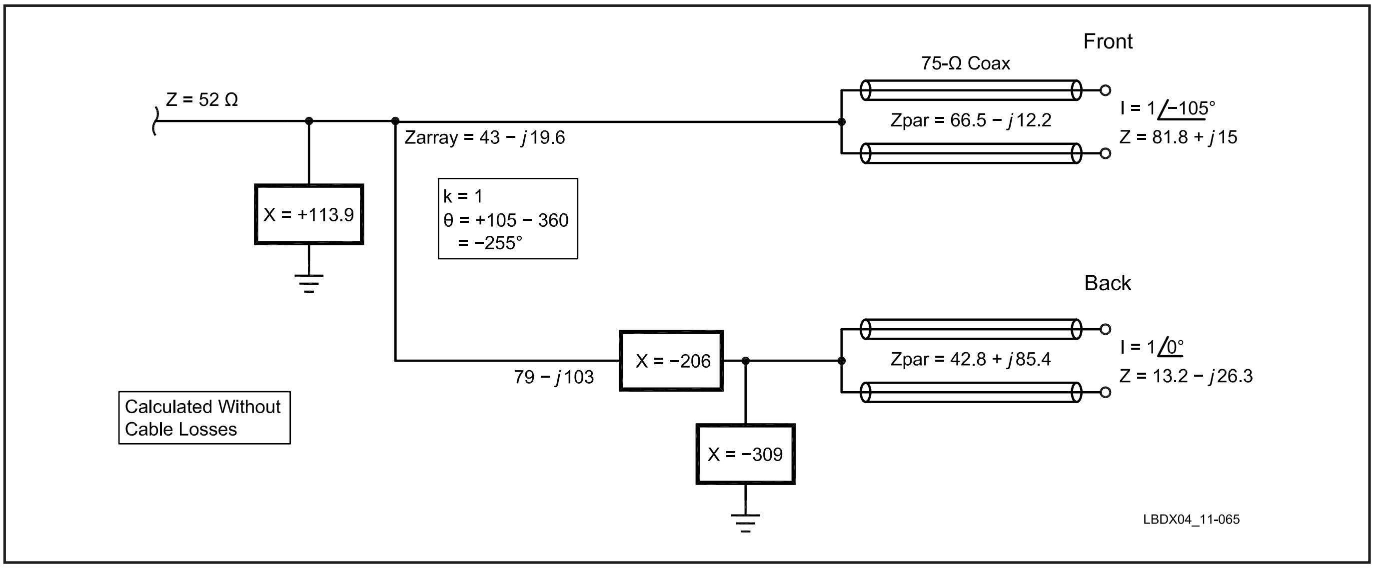
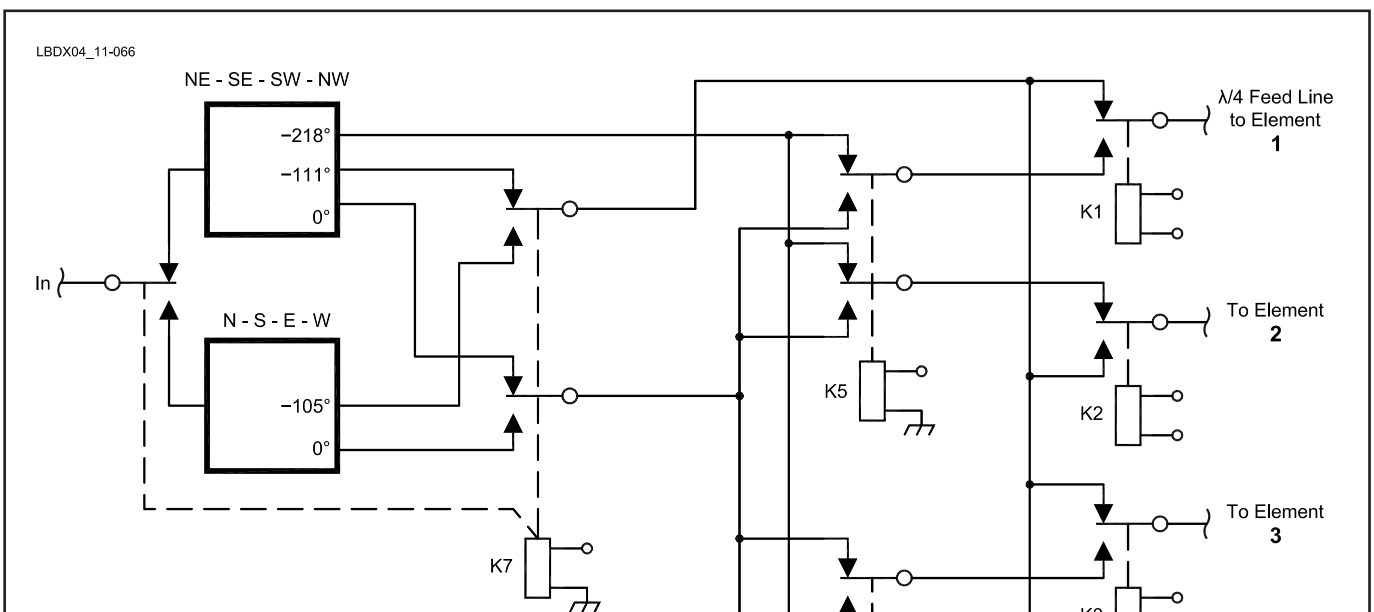



**Fig 11-65—Lahlum/Lewallen feed circuit for a Four-Square working as two closely spaced end-fire cells with optimized phasing, shooting along the directions of the side of the square.**




| | N | NE | E | SE | S | SW | W | NW |
|---|---|---|---|---|---|---|---|---|
| **K1** | X | O | O | X | O | O | 1 | 1 |
| **K2** | O | O | O | X | X | O | 1 | 1 |
| **K3** | O | O | 1 | X | X | O | O | 1 |
| **K4** | X | O | 1 | X | O | O | O | 1 |
| **K5** | O | O | O | X | O | 1 | O | O |
| **K6** | O | O | 1 | X | X | 1 | 1 | O |
| **K7** | X | O | X | O | X | O | X | O |

**Fig 11-66—Direction-switching for an optimized Four-Square (WA3FET ) in an intermediate direction. Two feed networks are used, selected by relay K7. The feed network for the main directions (NE, SE, SW and NW) is shown in Fig 11-65.**

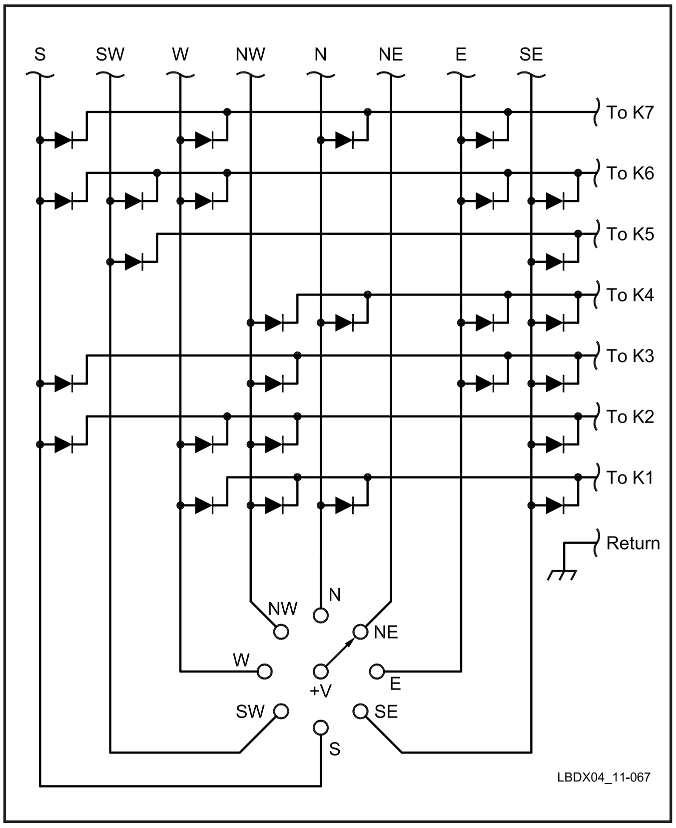



**Fig 11-67—Diode matrix for switching the Four-Square in eight directions. This circuit can be applied to both Figs 11-63 and 11-66. When used with Fig 11-63 you do not use the relay line "to K7".**

–61.4 Ω and Xs = +66.4 Ω), but the disadvantage is that the array impedance is relatively low at 16 + *j* 24 Ω.

### *4.8.2.3 Direction switching, optimized 8-direction Four-Square*

**Fig 11-66** shows a possible direction switching method, similar to the one used in Fig 11-63, with the difference that there is one extra relay (K7) for switching the two networks. In **Fig 11-67** we see the truth table from Fig 11-66 translated to a diode-matrix switching system.

### 4.8.3. Conclusion, 8-direction Four-Square

When you are happy with the quadrature-fed Four-Square, adding the intermediate directions is only a question of a slightly more complicated direction-switching system. As the hybrid coupler will see vastly different impedances when switching from the "full" directions to the "half" directions, the input impedance of the network will also be different, making fast switching impossible, unless you add two L-networks, which would be switched automatically. With the optimized versions things become a bit more complicated as you will require two different feed/phasing circuits.

It is questionable if all these efforts are worthwhile. Having been a user of a Four-Square with just four directions for over 10 years now, I have never felt the urge of adding four more directions. Take that for whatever you think it is worth!

## 4.9. The Broadside/End-Fire Array

This array has a spacing in the end-fire cells of 45 meters (on 160 meters) and a broadside spacing of 90 meters. The broadside spacing should be a minimum of 80 meters and can

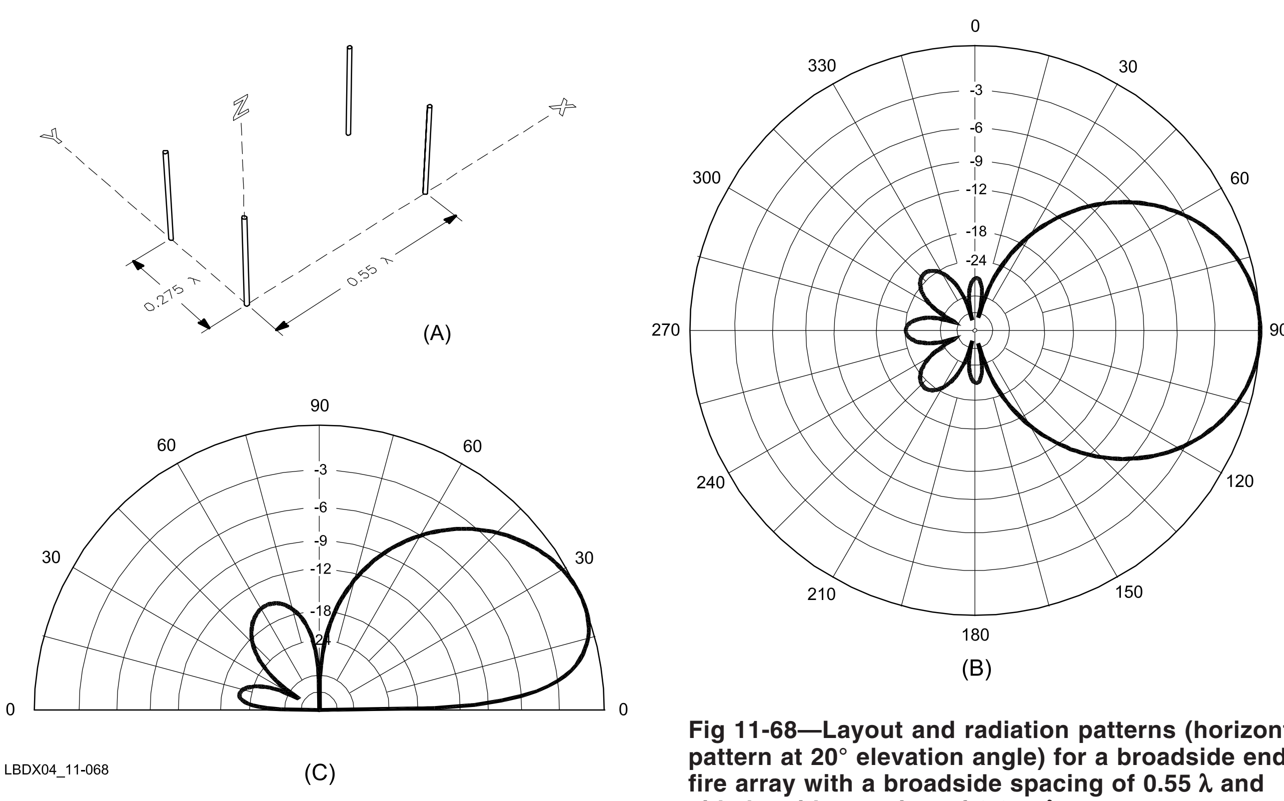



**Fig 11-68—Layout and radiation patterns (horizontal pattern at 20° elevation angle) for a broadside end-fire array with a broadside spacing of 0.55 λ and side-by-side spacing of 0.275 λ.**

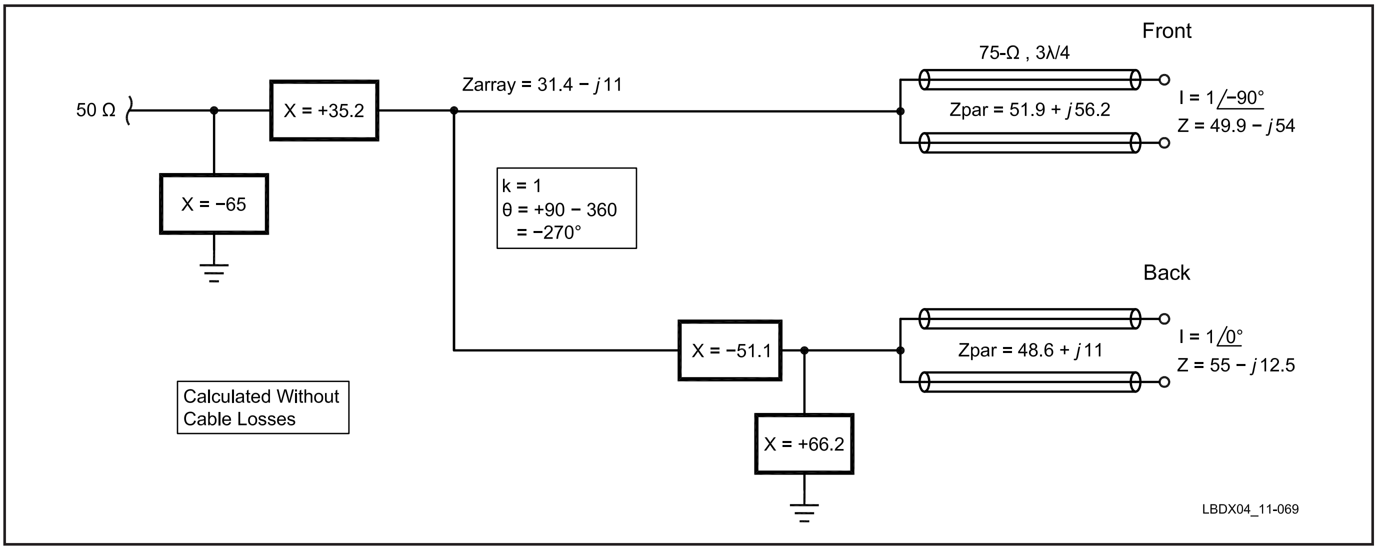



**Fig 11-69—Lewallen feed system (L-network) for the array in Fig 11-68. An L-network is included to match the feed system input impedance to 50 Ω.The alternative, where we feed the back elements directly and the front elements via an L-network, results in a much lower feed impedance (19 + *j* 12 Ω).**

be as much as 125 meters, achieving better rejection at high elevation angles but poorer directivity at low angles.

The end-fire cells could also have much smaller spacing, in which case a different phasing would be required; for example, 135° to 145° for λ/4 spacing. The bandwidth would however suffer from the small spacing. The array that was calculated is fed with 90° phase shift. These end-fire broadside combinations are also used in the 8-Circle array (Section 4.14).

### 4.9.1. Data, broadside/end-fire array

Broadside spacing: 0.55 λ
End-fire cell separation: 0.275 λ
End-fire cell phase: 90°
Feed currents:
Element 1 and Element 2 (back): I = 1 /0° A
Element 3 and Element 4 (front): 1 /–90° A
Gain = 8.43 dBi
3-dB forward angle: 57.4°
RDF = 12.11 dB
DMF = 20.8 dB
Feed impedances:
Z(Element 1) = Z(Element 2) = 55 – *j* 12.5 Ω (back)
Z(Element 3) = Z(Element 4) = 38.6 + *j* 11 Ω (front)

### 4.9.2. Feed system, broadside/end-fire array

This array, which is quadrature-fed (in 90° increments) can use a hybrid coupler. I designed a Lewallen LC-network type coupler, which has the advantage of being able to tune the array. The network was designed around 75-Ω current-forcing feed lines, which results in higher impedances than when using 50-Ω lines. Note that you need to use 270° long feed lines because of the physical separation of the two end-fire cells. **Fig 11-69** shows the Lewallen feed system for the array.

Direction switching is very simple, all you need is a single DPDT relay to invert the paired feed lines to the front and to back (points indicated as Fr and Bk in Fig 11-69).

## 4.10. The 5-Rectangle Array

One way of getting radiation "off the side" from an end-fire broadside array (see Section 4.5) with a decent pattern is by adding a fifth radiator right in the center of the rectangle. **Fig 11-70** shows the configuration and the radiation patterns obtained.

### 4.10.1. Data, 5-Rectangle array

Length of rectangle: 0.55 λ
Width of rectangle: 0.275 λ
Feed currents: Element 1 and Element 2 (back):
I = 1/0° A
Element 5 (center): I = 3 /–90° A
Element 3 and Element 4 (front): I = 1 /–180° A
Gain = 5.92 dBi
3-dB forward angle: 122°
RDF = 9.82 dB
DMF = 19.4 dB
Feed impedances:
Z (Element 1) = Z (Element 2) = 23.3 – *j* 37.3 Ω (back)
Z (Element 5) = 35.4 + *j* 1.8 Ω (center, middle)
Z (Element 3) = Z (Element 4) = 130 + *j* 30 Ω (front)

### 4.10.2 Feed system, 5-Rectangle array

The Lahlum/Lewallen LC-network feed system shown in **Fig 11-71** feeds the central element directly. The front and back elements are fed via LC networks, the front element with a –90° phase shift, the back element with a phase shift of +90°, which equals –270°.

### 4.10.3. Direction switching, 5-Rectangle array

As the 5-Rectange is a configuration that will only be used together with an end-fore/broadside array for giving right-angle coverage, I designed a direction-switching system that switches both configurations. In **Fig 11-72** North and South are the high-gain directions, and East and West the "fill-in" directions.

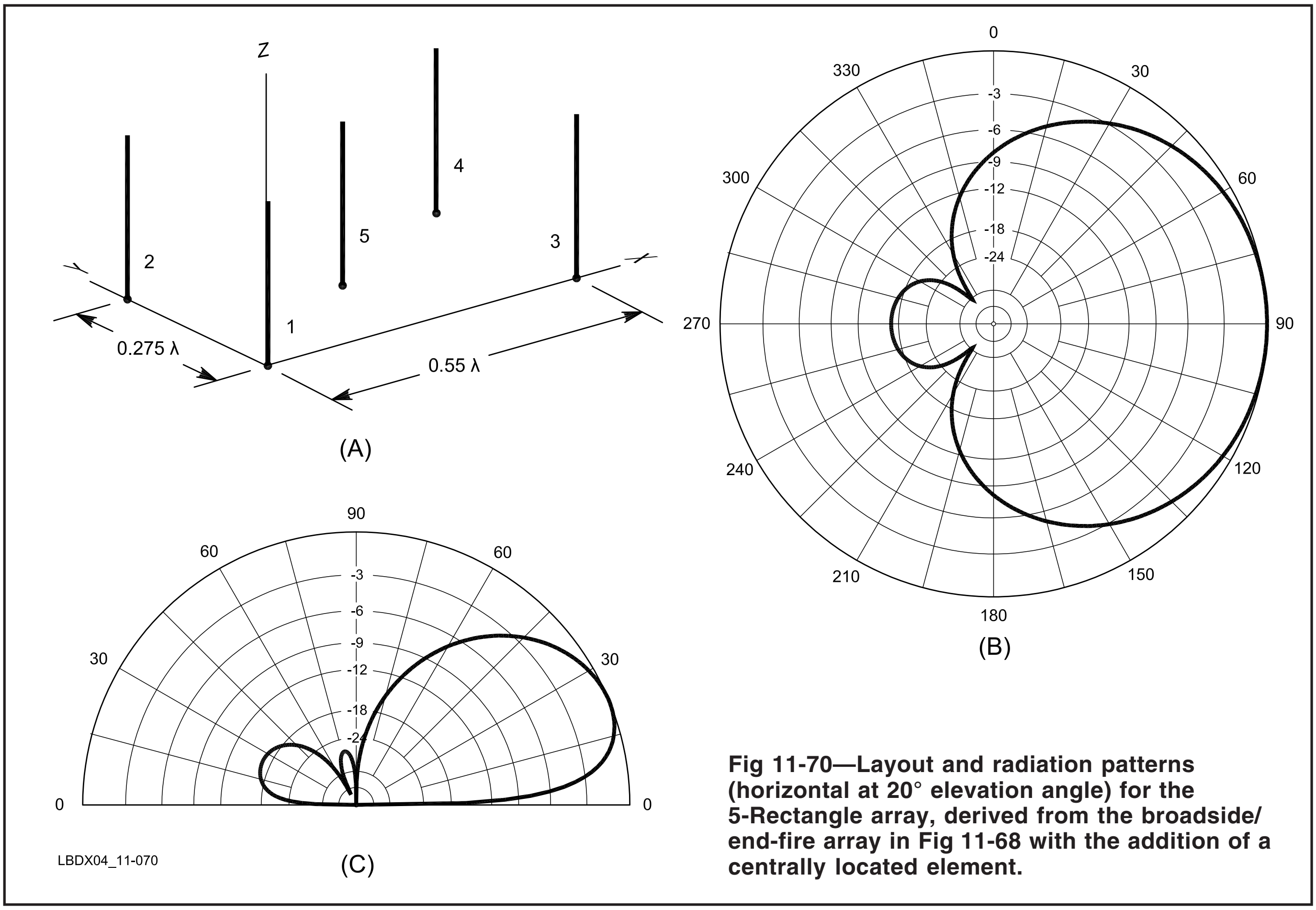



**Fig 11-70—Layout and radiation patterns (horizontal at 20° elevation angle) for the 5-Rectangle array, derived from the broadside/end-fire array in Fig 11-68 with the addition of a centrally located element.**

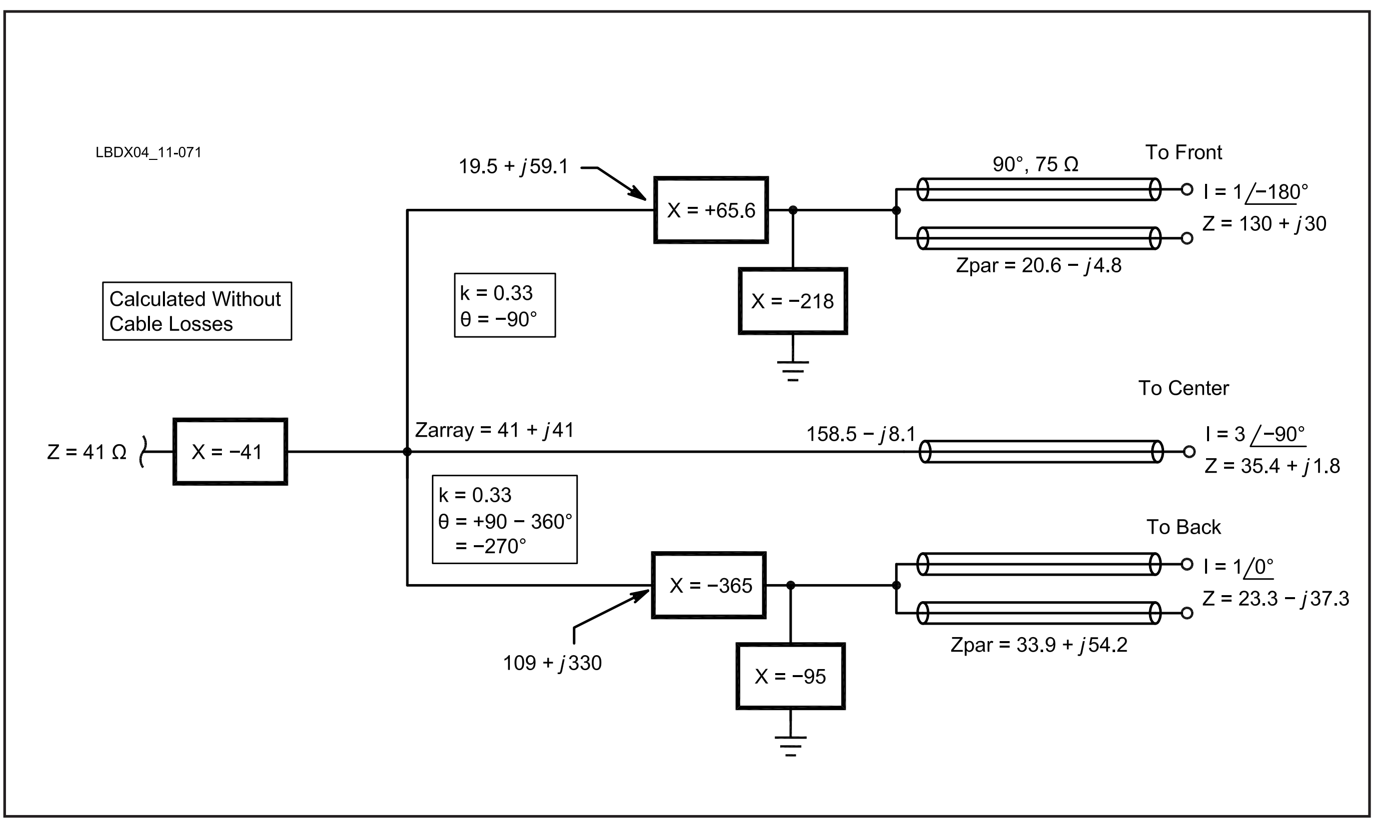
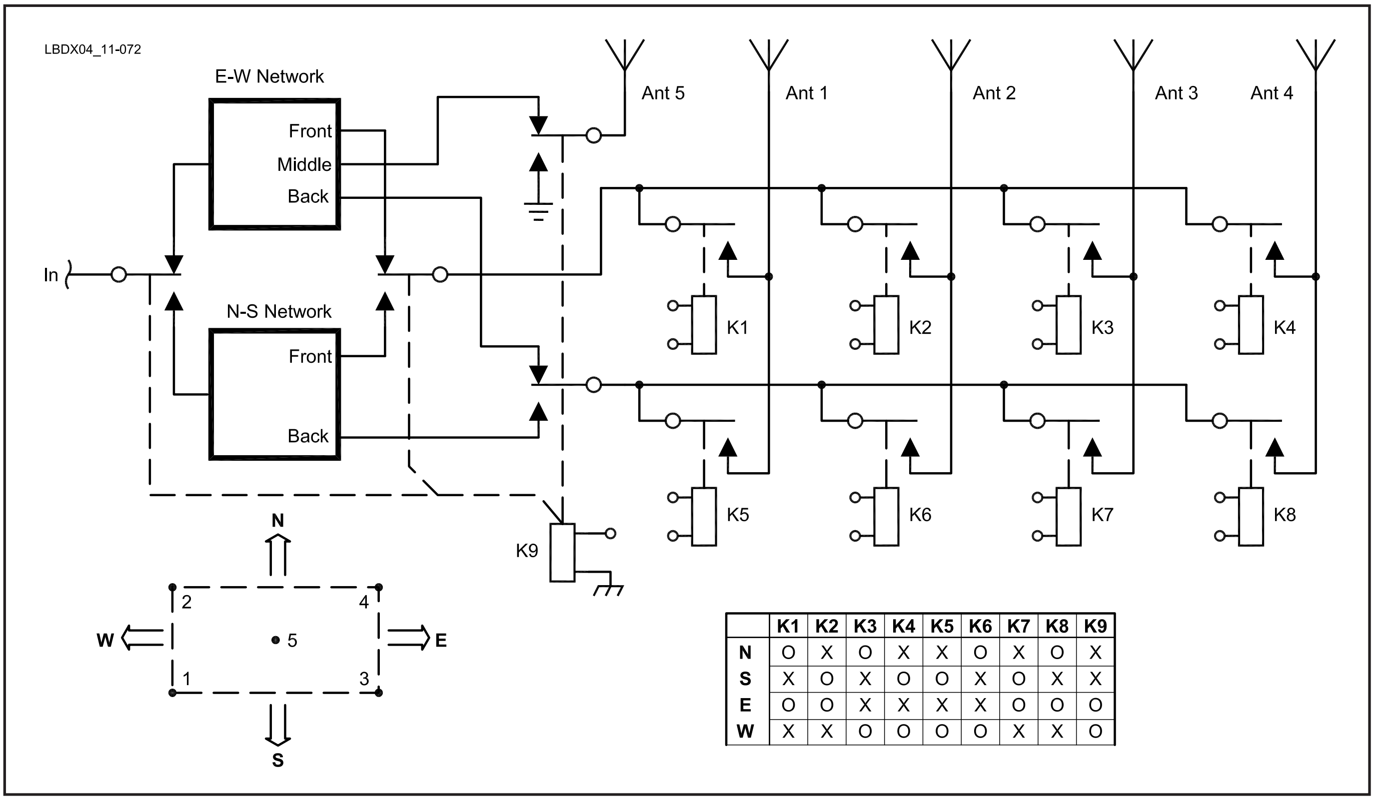



**Fig 11-71—Lahlum/Lewallen Feed system for the 5-Rectangle.**




| | K1 | K2 | K3 | K4 | K5 | K6 | K7 | K8 | K9 |
|---|---|---|---|---|---|---|---|---|---|
| N | O | X | O | X | X | O | X | O | X |
| S | X | O | X | O | O | X | O | X | X |
| E | O | O | X | X | X | X | O | O | O |
| W | X | X | O | O | O | O | X | X | O |

**Fig 11-72—Direction-switching for the 5-Rectangle, as fill-in directions for a broadside/end-fire array described in Section 4.7.**

**Fig 11-73—Russian-made small vacuum relays having single make/break contacts. These are well-suited for a relay matrix.**

You will need to build two networks and the right feed network will be selected by K9. The selection of the elements utilizes a small relay matrix consisting of eight relays. **Fig 11-73** shows a Russian-made vacuum relay that I use for switching.

## 4.11. Five-Square Array

The 5-Square is a modified Four-Square as shown in **Fig 11-74.** The side of square is 0.3 λ. The array can be made to cover eight directions. Shooting along the X and the Y axis the array has one reflector (Element 5), one director (Element 4) and three elements (1, 2 and 3) fed in phase, although with slightly different current magnitudes.

### 4.11.1. Data, "diagonal" operation of Five-Square array

Gain = 7.67 dBi (and that is 1 dB better than a classic Four-Square)
3-dB forward angle: 78°
RDF = 11.92 dB
DMF = 23.0 dB
Feed currents:
Element 5 (back): I = 1.25 /0° A
Element 1 (center): I = 1 /–125° A
Element 2 = Element 3 (outside center): 1 /–125° A
Element 4 (front): I = 1.25 /–255° A
Feed impedances:
Z (Element 5) = 2.5 – *j* 2.9 Ω (back)
Z (Element 1) = 45.4 – *j* 6.5 Ω (center, middle)
Z (Element 2) = Z (Element 3) = 37.2 – *j* 1.7 Ω (outside center)

Z(Element 4) = 28.5 + *j* 61 Ω (front)

Using the same physical layout we can shoot along the bisector of the X-Y axis, adding another four directions to the array. In this configuration we have two directors (Elements 4 and 2) and 2 reflectors (Elements 3 and 5).

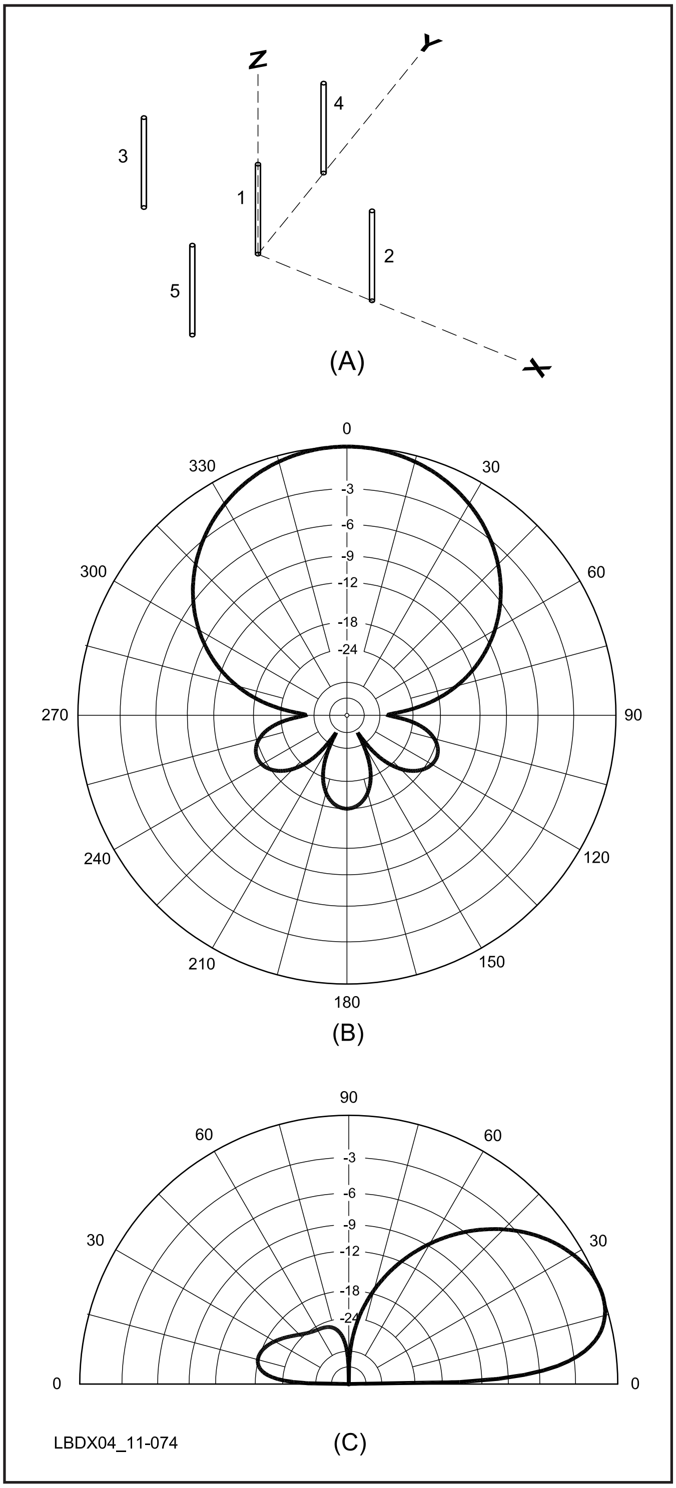



**Fig 11-74—Layout and radiation patterns (horizontal at 20° elevation angle) for the Five-Square array operating in the diagonal fashion. The excellent directivity is mainly derived from its relatively narrow forward lobe (typically 20-25° less than for a Four-Square).**

### 4.11.2 Data, half-way angles of Five-Square array

Gain = 6.2 dBi
3-dB forward angle: 102.4°
RDF = 10.52 dB
DMF = 18.5 dB
Feed currents:
Element 3 = Element 5 (back): I = 0.6 /0° A
Element 1 (center): I = 2 /–110° A
Element 4 = Element 2 (front): I = 0.7 /–230° A
Feed impedances:
Z (Element 3) = Z (Element 5) = 8.8 – *j* 32.5 Ω (back)
Z (Element 1) = 22.8 + *j* 5.1 Ω
Z (Element 4) = Z (Element 2) = 23.4 + *j* 68.1 Ω (front)

In this configuration the performance is substantially the same as obtained by a Four-Square à la WA3FET. You can get a better F/B by changing the feed currents but will have a much

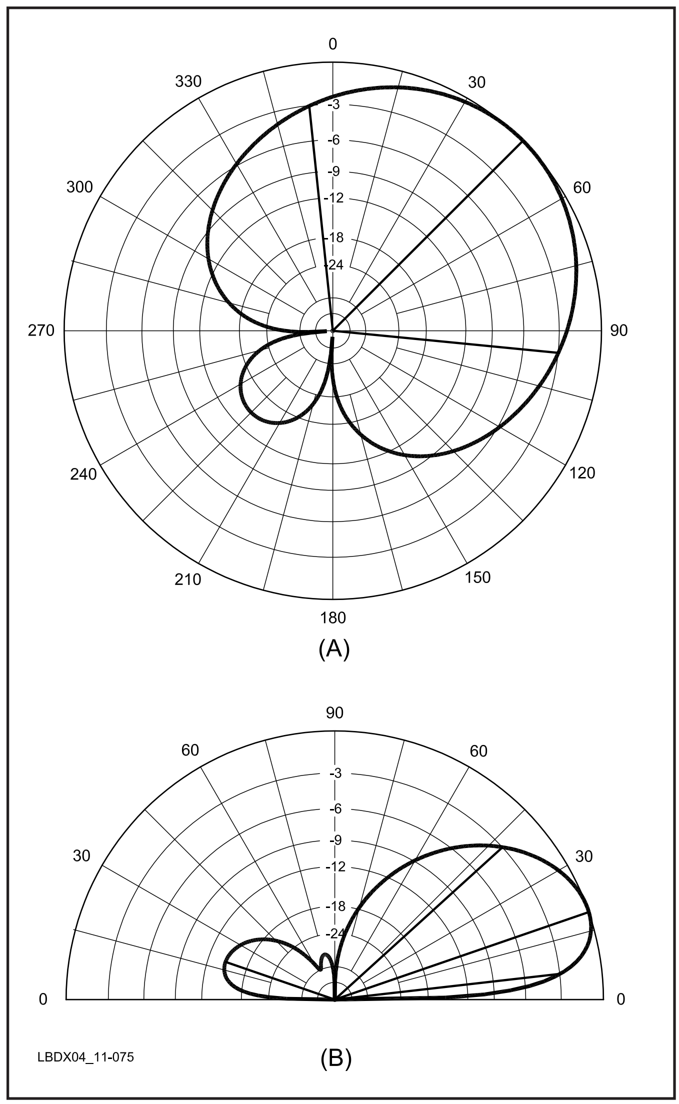



**Fig 11-75—When shooting along the sides of the square, the array operates with one center element with two directors and two reflectors. The 3-dB forward lobe angle is roughly the same as for a classic Four-Square, and its gain is similar to a Four-Square optimized by WA3FET.**

wider forward lobe, decreasing the RDF. By playing with a modeling program you can fine-tune this array to your liking.

It is obvious that the "main" directions are those along the X and Y-axes with 1.6 dB more gain and 1.4 dB more RDF (directivity), mainly obtained through a substantially narrower 3-dB forward beamwidth (78° vs 102°). In both configurations the array can be fed using the Lahlum/Lewallen feed system.

A nice thing about this Five-Square is that you don't really need five towers—You can hang the center element from some nylon or Dacron catenary cables strung between four towers. If the center element is slightly shorter because of the sag of the catenary cables, just top load it with four cross wires, running in the direction of the support cables.

### 4.11.3. Feed system, Five-Square array

#### *4.11.3.1. Diagonal operation, Five-Square array*

**Fig 11-76** and **Fig 11-77** show two feed systems developed according the Lahlum/Lewallen system, using the "Lahlum.xls" spreadsheet. If you want to make the array cover eight directions, you will have to install both networks, and switch them in and out of the circuit according to the direction used.

### 4.11.4. Switching directions, Five-Square array

**Fig 11-78** shows the direction switching, accomplished

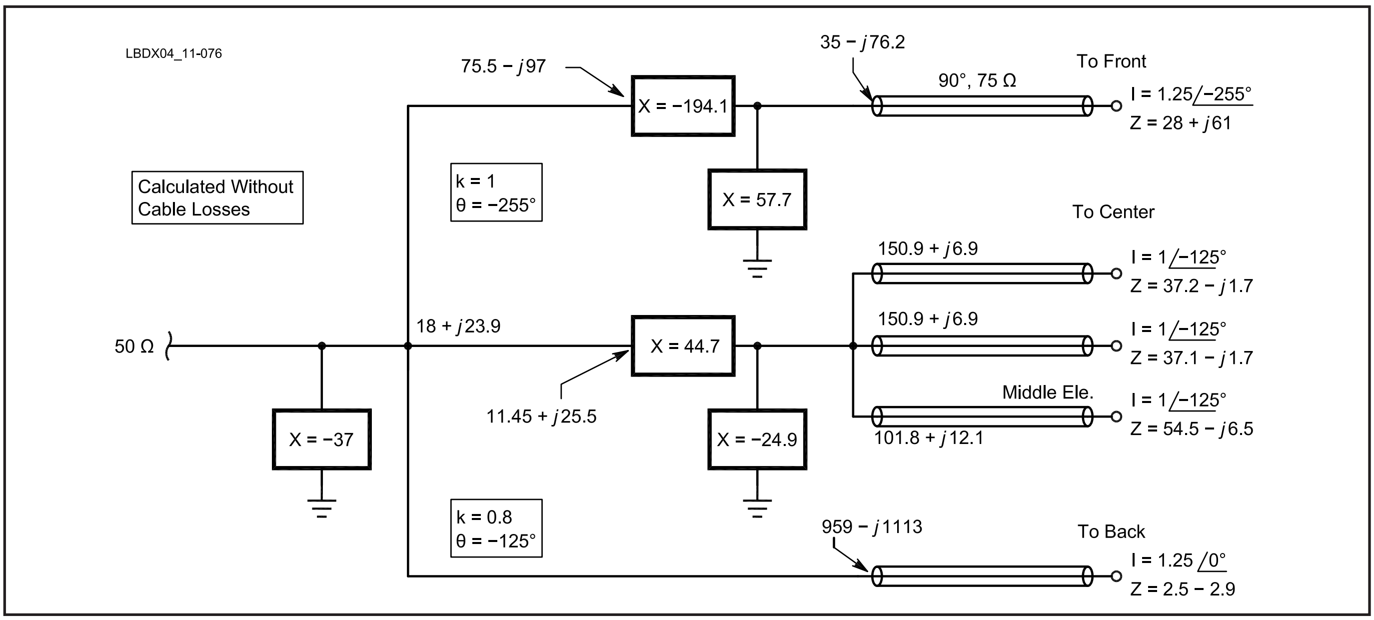



**Fig 11-76—Lahlum feed network for the Five-Square array shooting diagonally across the square in one of its main directions.**

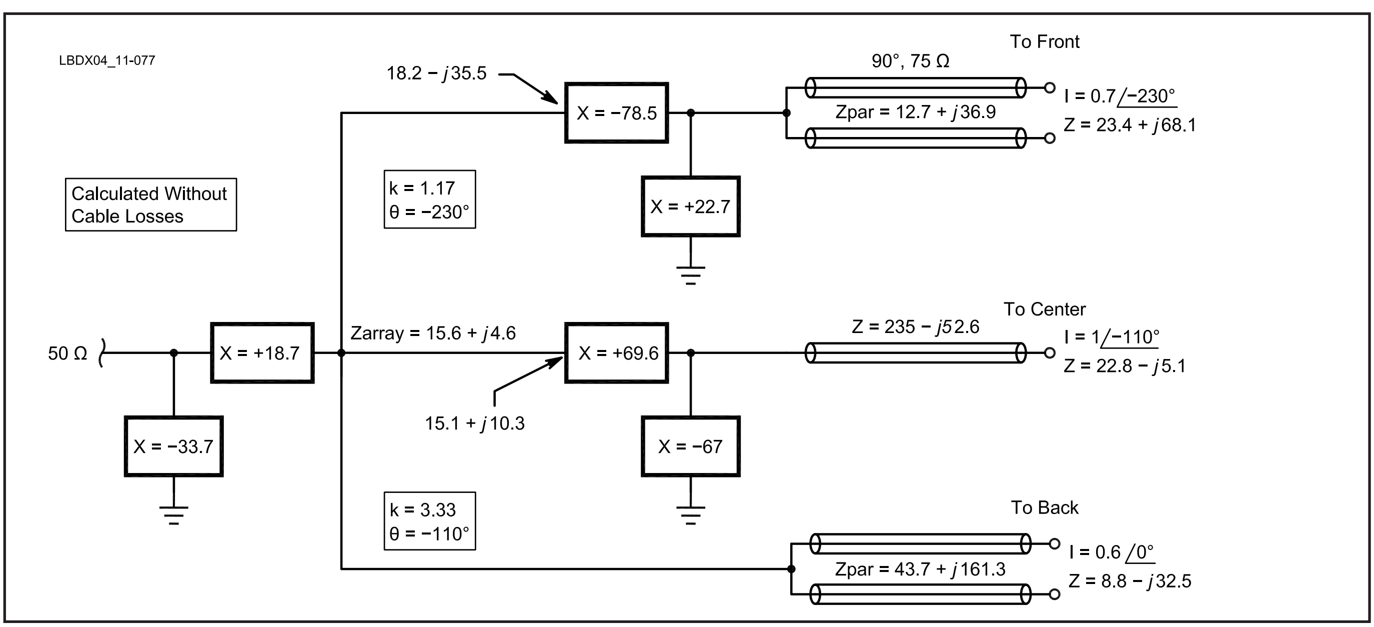



**Fig 11-77—Lahlum feed network for the Five-Square array shooting along the directions of the sides of the square in the "intermediate" directions.**

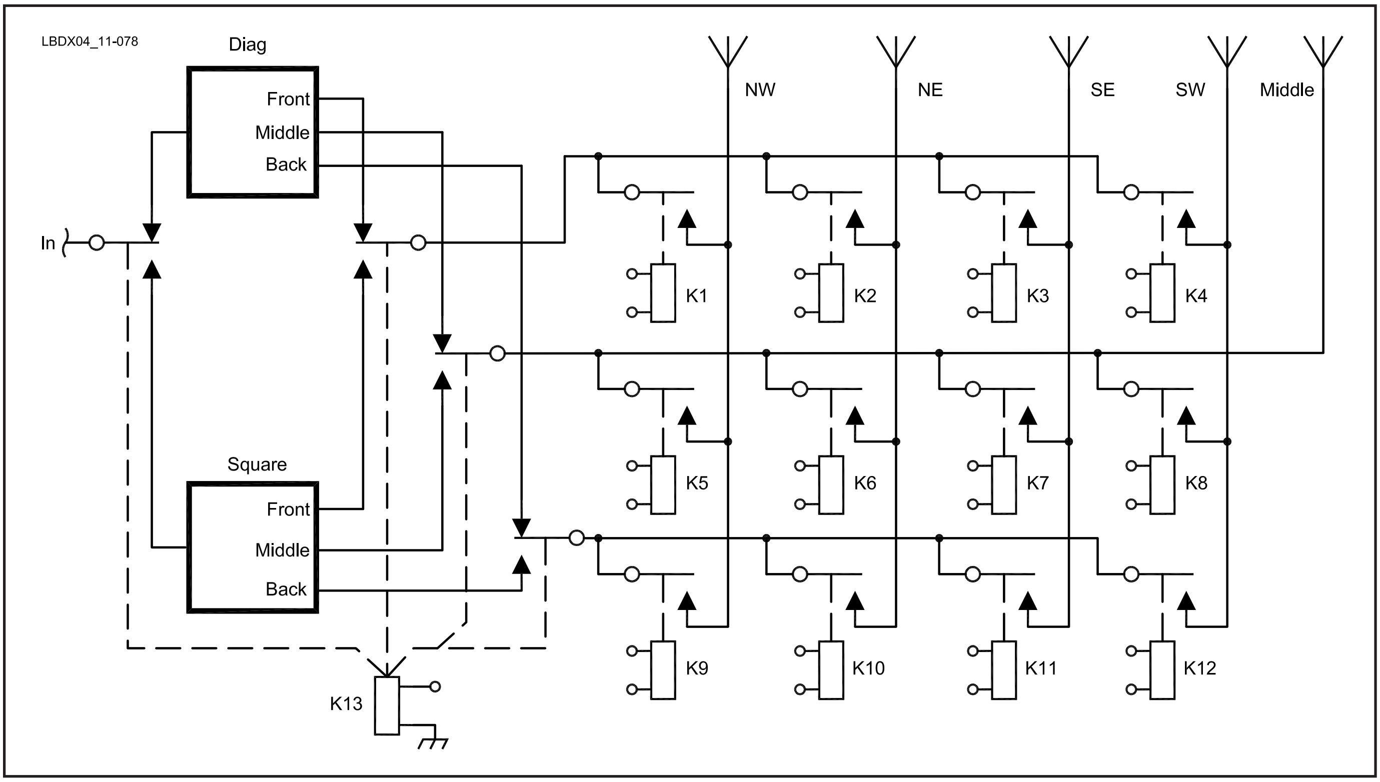



**Fig 11-78—Direction-switching system for the Five-Square array.**

**Table 11-14**
**Truth Table of Relay Matrix for Five-Square Array**

| | *K1* | *K2* | *K3* | *K4* | *K5* | *K6* | *K7* | *K8* | *K9* | *K10* | *K11* | *K12* | *K13* |
|---|---|---|---|---|---|---|---|---|---|---|---|---|---|
| N | X | X | O | O | O | O | O | O | O | O | X | X | X |
| NE | O | X | O | O | X | O | O | X | O | O | O | X | O |
| E | O | X | X | O | O | O | O | O | X | O | O | X | X |
| SE | O | O | X | O | O | X | O | X | X | O | O | O | O |
| S | O | O | X | X | O | O | O | O | X | X | O | O | X |
| SW | O | O | O | X | X | O | X | O | O | X | O | O | O |
| W | X | O | O | X | O | O | O | O | O | X | X | O | X |
| NW | X | O | O | O | O | X | O | X | O | O | X | O | O |

with a small relay matrix. According to the selected direction the appropriate drive network is selected with relay K13. **Table 11-14** below shows the relay truth table.

### 4.11.5. Conclusion, Five-Square array

For 160 meters, and including 40-meter-long radials, this array requires an area measuring 107 by 107 meters, (1.15 hectares, or ~ 3 acres). With 40-meter long radials, the antenna has a foot print that is only 20% larger than for the classic Four-Square (with $\lambda/4$ sides), yet produces 2 dB more gain and has the possibility of switching in eight directions. Clearly a winner! If you care for the eight directions you will, of course, need two different sets of L-networks to establish the required phased shifts.

With 120 radials that are 20 meters long for all elements, the footprint is 0.8 hectares (approximately 2 acres), and the trade for gain will be marginal (a fraction of a dB). Obviously for 80 meters, the required real estate is four times smaller. **Fig 11-79** is a photograph of NO8D's Five-Square array.

## 4.12. The 6-Circle Array

The 6-Circle was described in Chapter 7 as a receiving antenna. I developed two transmit versions, one having a circle diameter of 80 meters and one smaller version with 60-meter diameter (for 1.83 MHz). Still smaller versions (eg, 30-meters diameter) work very well as receiving arrays, but have low feed impedances and narrow bandwidth when used as transmit antennas.

### 4.12.1. Array data (40-meter radius), 6-Circle array

Circle diameter: 80 meters
Frequency: 1.83 MHz
Gain = 7.8 dBi
3-dB forwards angle: 74.6°

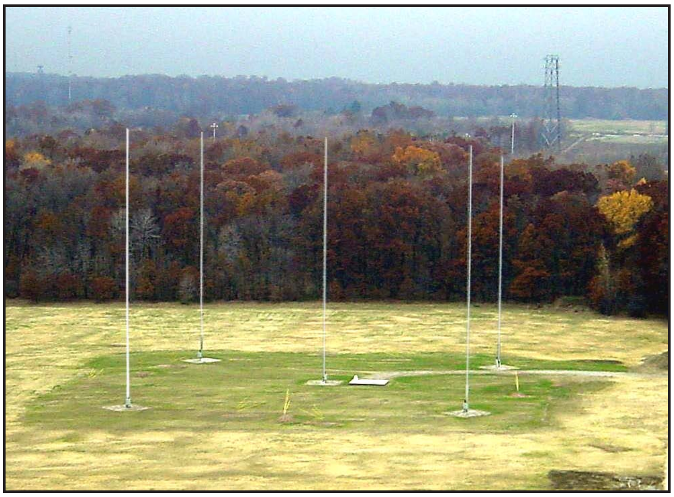


Fig 11-79—160-meter Five-Square at NO8D.

RDF = 11.63 dB
DMF = 24.2 dB
Feed currents:
Element 5 and Element 6 (back): 1 /0° A
Element 1 and Element 2 (center): I = 2 /–90° A
Element 3 and Element 4 (front): I = 1 /–180° A
Feed impedances:
Z (Element 5) = Z (Element 6) = 5.3 – *j* 15.4 Ω (back)
Z (Element 1) = Z (Element 2) = 27.9 – *j* 11.6 Ω
Z (Element 4) = Z (Element 3) = 129 + *j* 48.8 Ω (front)

This version with a circle radius of 40 meters is quadrature fed. We can feed it with a hybrid coupler (see Section 3.4.6) but since the feed current in the center elements is twice as high as for the other elements we need to feed them with two paralleled feed lines.

If you decide to feed this array using a hybrid coupler, use 3λ/4-long feed lines of 75-Ω rather than 50-Ω impedance, which will result in less power lost in the hybrid's dummy load. One issue with an 80-meter diameter 6-Circle array is that λ/4-long feed lines do not reach the center of the array. This requires 3λ/4 lines, with all the attendant drawbacks (more loss, more cost and much less bandwidth).

For this reason I developed a slightly smaller 6-Circle, which has a radius of 30 meters. This accommodates λ/4 feed lines using foam coaxial cables with VF of approx 0.8.

### 4.11.2 Array data (30-meter radius), 6-Circle array

Circle diameter: 60 meters
Frequency: 1.83 MHz
Feed currents:
Element 5 and Element 6 (back): 1 /0° A
Element 1 and Element 2 (center): I = 2 /–110° A
Element 3 and Element 4(front): I = 1 /–225° A
Gain = 7.6 dBi
3-dB forward angle: 78.2°
RDF = 11.72 dB
DMF = 25.5 dB
Feed impedances:
Z (Element 5) = Z (Element 6) = 4.3 – *j* 16.1 Ω (back)
Z (Element 1) = Z (Element 2) = 25.2 – *j* 7.7 Ω
Z (Element 4) = Z (Element 3) = 52 + *j* 93 Ω (front)

This smaller version of the 6-Circle is no longer quadra-

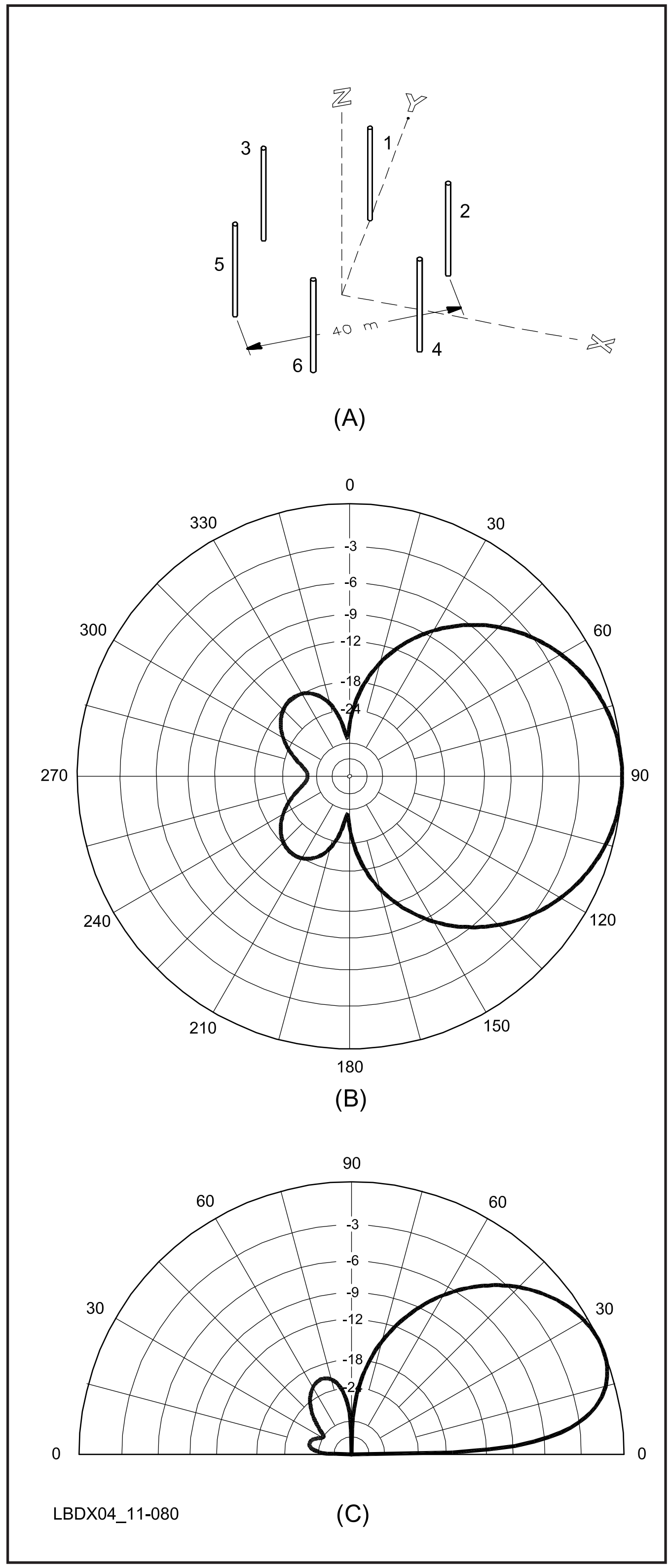



Fig 11-80—Layout and radiation patterns (horizontal pattern at 20° elevation angle) for the 6-Circle array at 3.7 MHz. The patterns remain almost identical for different sizes of the array. Larger arrays will show higher impedances and will be somewhat easier to feed. They also have somewhat more bandwidth. Larger arrays will require 3λ/4 current-forcing feed lines, which have definite disadvantages when it comes to bandwidth. If you have a diameter of 30 meters, λ/4 feed lines will reach if you use foam coax with Vf ~ 0.8.

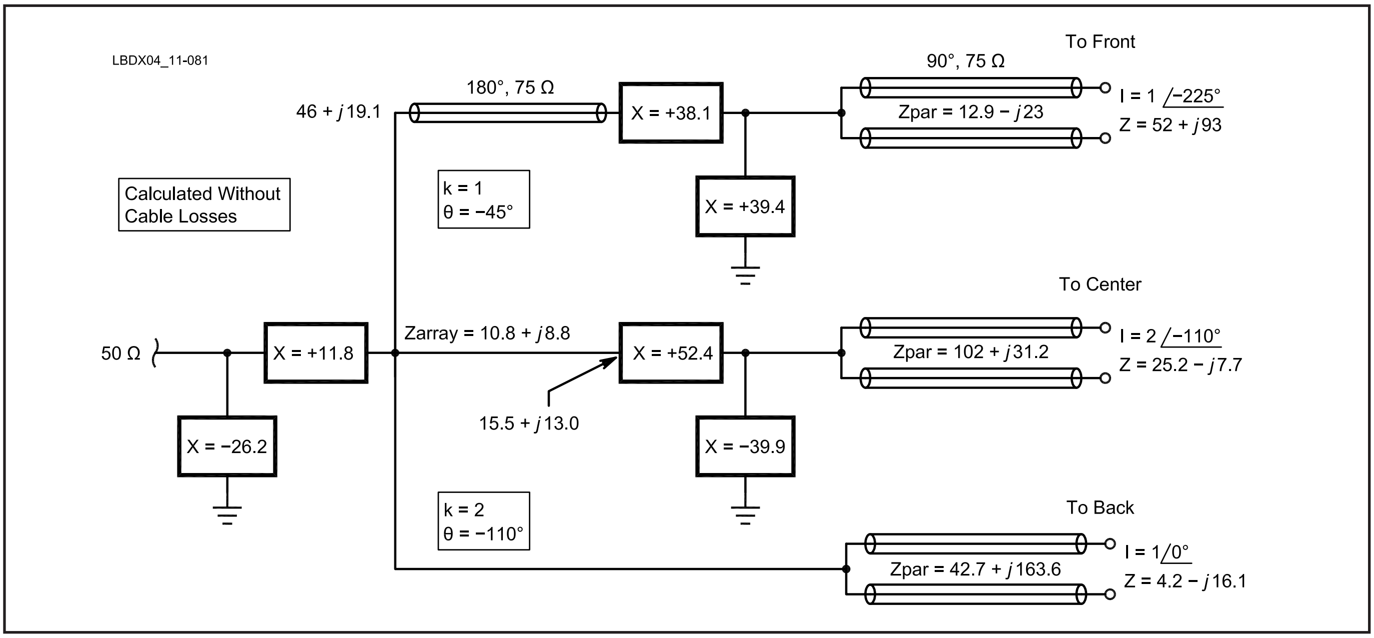
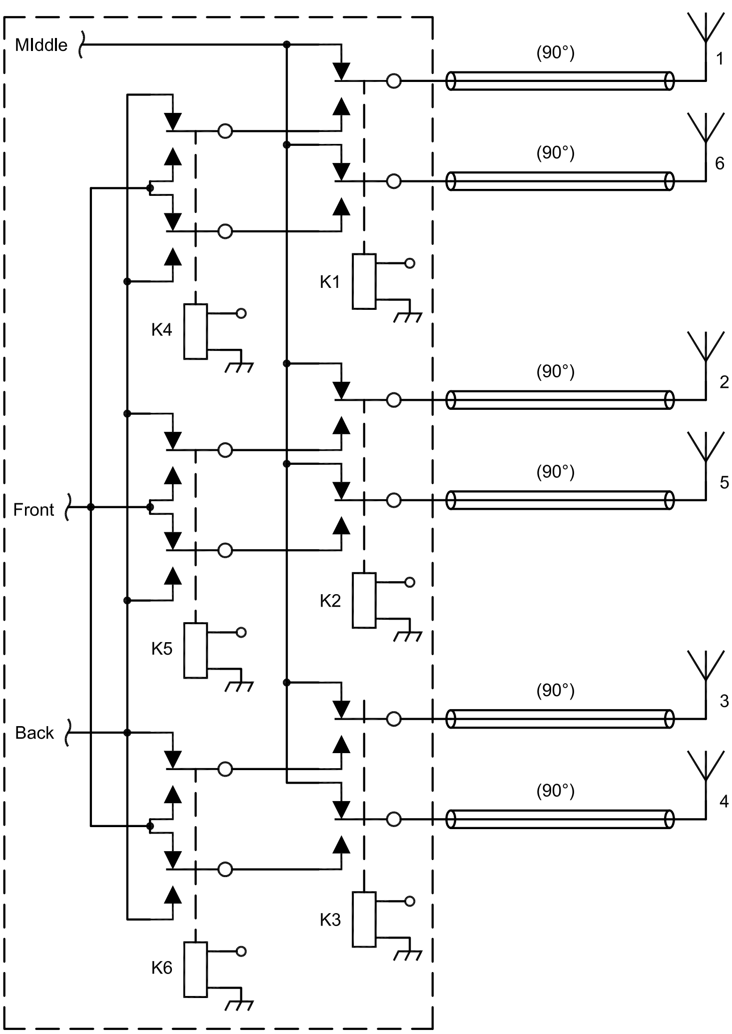
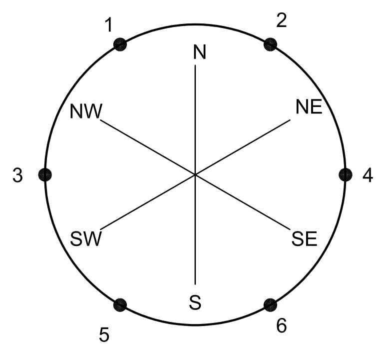



**Fig 11-81—Feed system for a "small" 6-Circle (radius = 30 meters). In this array the back elements are fed directly, the center elements via an L-network and the front elements via a 180° line plus L-network. This results in the most realizable network-components reactances, but a low feed impedance. Other solutions result in a higher feed impedance but require high L-network component reactances.**




| | N | NW | SW | S | SE | NE |
|---|---|---|---|---|---|---|
| **K1** | X | X | O | X | X | O |
| **K2** | X | O | X | X | O | X |
| **K3** | O | X | X | O | X | X |
| **K4** | X | X | - | O | O | - |
| **K5** | X | - | O | O | - | X |
| **K6** | - | O | O | - | X | X |

**Fig 11-82—Direction switching for the 6-Circle array. The lines "Middle" (middle elements). "Front" and "Back" connect to the points marked Center, Front and Back in Fig 11-81.**

LBDX04_11-082

ture-fed and requires the Lewallen/Lahlum feed system. Using this feed system there is no need for parallel feed lines, since the double feed current magnitude can be obtained through a proper L-network circuit (see Section 3.4.5).

### 4.12.3. Feed system, 6-Circle array

The Lahlum/Lewallen feed system in **Fig 11-81** feeds this smaller array. Here too, 75-Ω current-forcing feed lines are used to achieve a higher input impedance (which is only about 16 Ω if using 50-Ω feed lines).

### 4.12.4. Direction switching, 6-Circle array

**Fig 11-82** shows the direction switching circuitry for the 6-Circle array. Six DPDT relays are required.

### 4.12.5. Conclusion, 6-Circle array

On 160 meters, using 40-meter long radials, this array requires an area measuring 140 by 140 meter, approximately 2 hectares or 5 acres. It is a very nice performer, with excellent directivity and it should appeal to all those who want a top-notch array but just do not have the real estate for a 9-Circle (see Section 4.15).

With 120 radials, each 20-meters long, on each of the verticals the foot print is reduced to only 1 hectare (2.5 acres), with a loss in gain of just a fraction of a dB.

## 4.13. The 7-Circle Array

I developed a 7-Circle array using the same 60-meter circle diameter. This makes it possible to use λ/4 feed lines (with foam coax, Vf ~ 0.8) on 160 meters. This is identical to the 6-Circle but has another element in the center of the circle. It did not seem to be possible to improve the performance of the 6-Circle though.

### 4.13.1. Data, 7-Circle array

Gain = 7.61 dBi
3-dB forward angle: 80.4°
RDF = 11.74 dB
DMF = 26.3 dB
Feed currents:
Element 5 and Element 6 (Back): 0.8 /0° A
Element 1, Element 2, Element 7 (middle): I = 1 /–120° A
Element 3 and Element 4(front): I = 0.73 /–250° A
Feed impedances:
Z (Element 5) = Z (Element 6) = 11.5 – *j* 21.1 Ω (back)
Z (Element 1) = Z (Element 2) = 35.3 – *j* 2.6 Ω (outer middle elements)
Z (Element 7) = 40.5 – *j* 1.5 Ω (middle, center element)
Z (Element 4) = Z (Element 3) = 13 + *j* 88 Ω (front)

Feed system: The same schematic as used for the 6-Circle (Section 4.12) can be used, as the seventh element needs no switching when changing directions.

### 4.13.2. Conclusion, 7-Circle array

Nothing seems to be gained with the extra element in the middle.

## 4.14. The 8-Circle Array

In an 8-Circle array (see Chapter 8), we only use four elements at a time. The 8-Circle consists of two broadside cells, spaced about 0.65 λ, each cell consisting of a 2-element

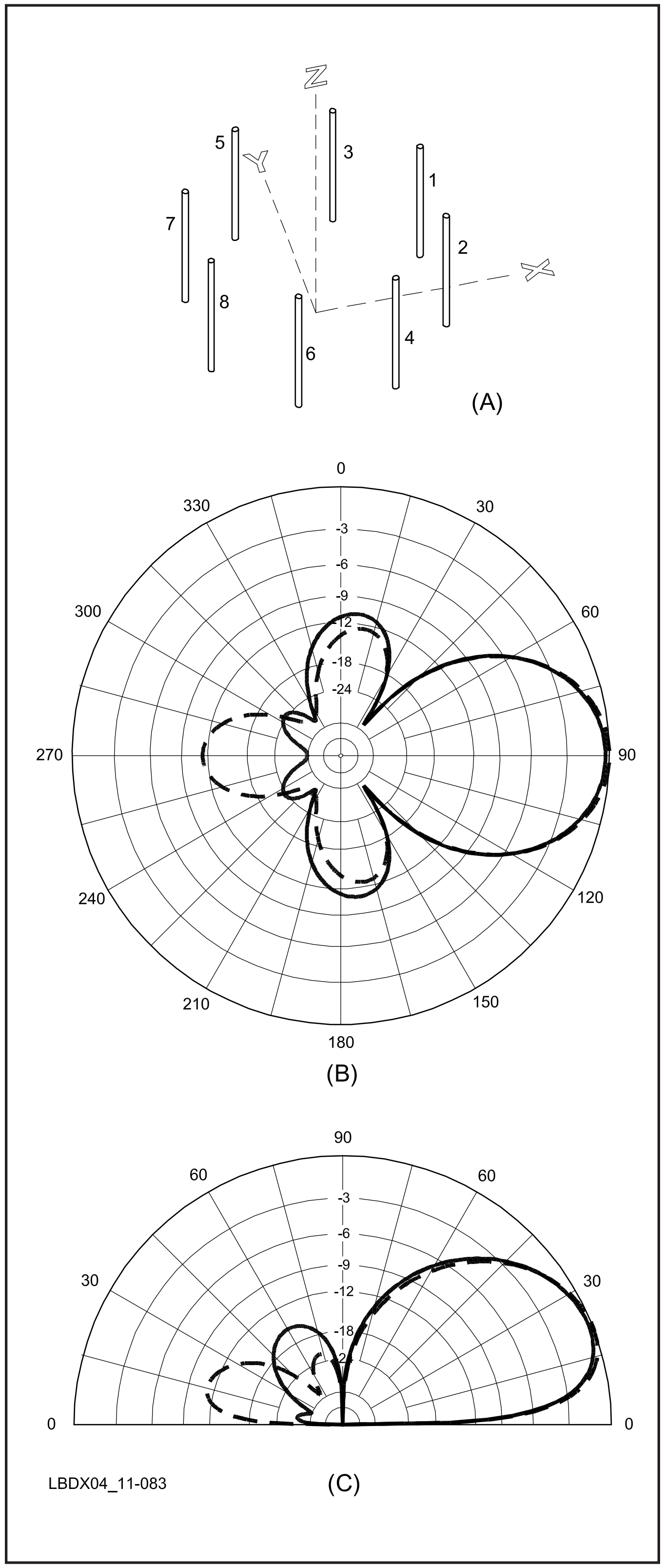



**Fig 11-83—At A, layout of the 8-Circle array. At B, comparison of azimuthal patterns at 20° elevation for phase shifts of 90° (solid line) and 120° (dashed line) in the end-fire cells. At B, elevation pattern comparison. Note good F/B at low angles for 90° phasing but large sidelobe at 60° elevation. For 120° phasing the F/B is better at high angles. The large azimuthal sidelobes for both phases are due to 0.65-λ spacing, which also gives the narrow forward lobe.**

end-fire array.

Using a side-by-side separation of 0.65 λ, you can obtain the narrowest possible forward beamwidth without creating excessive sidelobes. At 1.83 MHz, this results in a circle diameter of 115.8 meters and a separation between the two elements of the end-fire cell of 44.3 meters.

### 4.14.1. Data, quadrature-fed 8-Circle array

Gain = 8.95 dBi
3-dB forward angle: 47.2°
RDF = 12.86 dB
DMF: 21.8 dB
Feed currents:
Element 5 and Element 6 (back): 1 /0° A
Element 3 and Element 4(front): I = 1 /–90° A
Feed impedances:
Z (Element 5) = Z (Element 6) = 12.1 – *j* 6.3 Ω (back)
Z (Element 4) = Z (Element 3) = 38.6 + *j* 5.6 Ω (front)

### 4.14.2. Optimized phasing, 8-Circle array

When feeding the cells with a larger phase angle (120° vs 90°) we can increase the gain and the RDF, although the F/B at low angles suffers. The data now are:
Gain = 9.2 dBi
3-dB forward angle: 46.0°
RDF = 13.29 dB
DMF = 21.8 dB
Feed currents:
Element 5 and Element 6 (Back): 1 /0° A
Element 3 and Element 4(front): I = 1 /–120° A
Feed impedances:
Z (Element 5) = Z (Element 6) = 10.8 + *j* 1.1 Ω (back)
Z (Element 4) = Z (Element 3) = 33.8 + *j* 11.5 Ω (front)

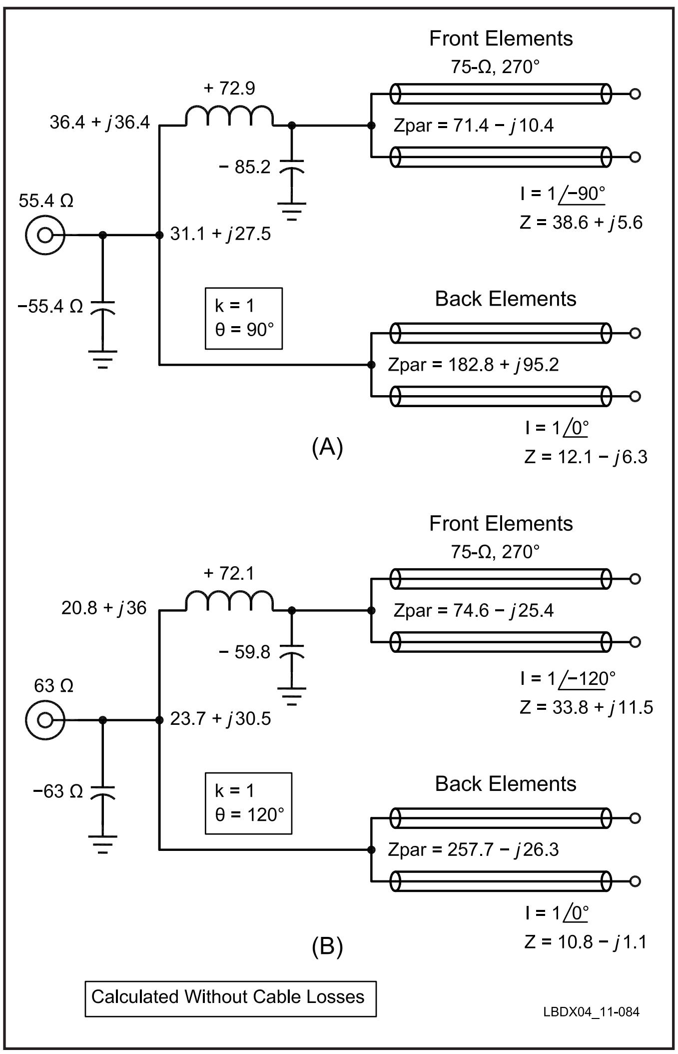



**Fig 11-84—Lahlum/Lewallen feed systems for the 8-Circle. At A, the values for a 90° phase shift. At B, the values when using a 120° phase shift.**

### 4.14.3. Feed system, 8-Circle array

When fed in quadrature (90° phase shift) the array can be fed using a hybrid coupler. Using a Lewallen L-network feed system gives the advantage of being able to adjust the network components to obtain the desired phase shift and current magnitude in the front element.

**Fig 11-84** shows the networks calculated with the *Lahlum.xls* spreadsheet. Both the 90° phase angle solution (A) and the 120° phase angle alternative (B) are shown. 75-Ω feed lines are recommended in order to achieve a high enough network input impedance. **Fig 11-85** shows a direction-switching system that can be used with the 8-Circle array.

### 4.14.4. Discussion, 8-Circle array

Of all large high-performance arrays, the 8-Circle is certainly the easiest one to build, as there are only two phase angles involved. The directivity is excellent, especially the RDF, the narrow forward lobe being responsible of this. With this array you will hardly need separate receiving antennas. Unfortunately, you will need at least 4 hectares (~ 10 acres) of real estate for this array on 160 meters, or 1 hectares (2.5 acres) on 80 meters!

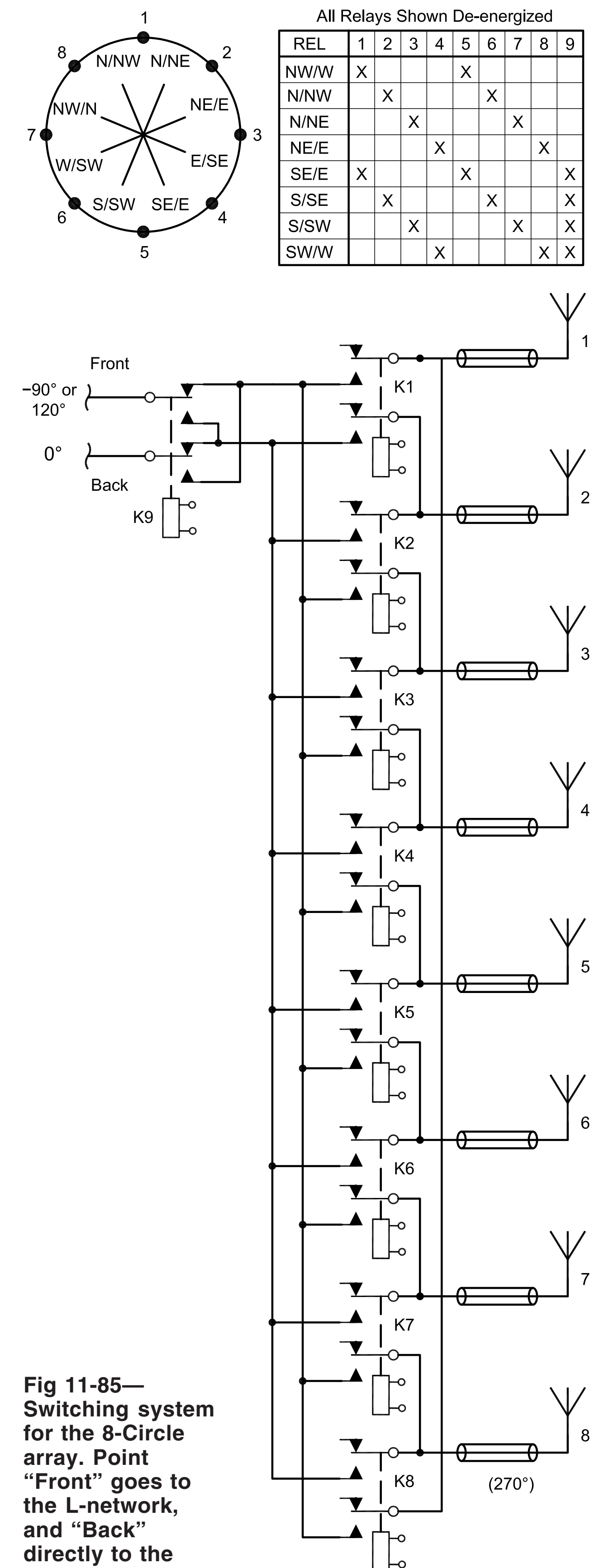
## 4.15. The 9-Circle Array

The 9-Circle was originally developed by John Brosnahan, WØUN. John Battin, K9DX, has a 160-meter array at his remote station near Chicago and is planning to build an 80-meter version during the summer of 2004. The other existing 80-meter 9-Circle is at the QTH of Paul Hellenberg, K4JA, in Virginia.

### 4.15.1 Data, 160-meter 9-Circle array

Size = 128 meters diameter
Gain = 9.05 dBi over Good Ground. (Over Very Good Ground (ε = 20 and conductivity = 30 mS), the gain is 11.0 dBi.)
3-dB forward angle: 58.5°
RDF = 12.99 dB
DMF = 31.7 dB
Feed currents:
Element 6 (back, tip): I = 1 /0° A
Element 5 and Element 7 (back, side-by-side): I = 1.66 /–90° A
Element 4 and Element 6 (middle, outer): I = 1 /–180° A
Element 1 (middle, center): I = 3 /–180° A
Element 3 and Element 9 (front, side-by-side): I = 1.66 /–270° A

Element 2 (front, tip): I = 1 /–360° A




All Relays Shown De-energized

| REL | 1 | 2 | 3 | 4 | 5 | 6 | 7 | 8 | 9 |
|---|---|---|---|---|---|---|---|---|---|
| NW/W | X | | | | X | | | | |
| N/NW | | X | | | | X | | | |
| N/NE | | | X | | | | X | | |
| NE/E | | | | X | | | | X | |
| SE/E | X | | | | X | | | | X |
| S/SE | | X | | | | X | | | X |
| S/SW | | | X | | | | X | | X |
| SW/W | | | | X | | | | X | X |

**Fig 11-85—Switching system for the 8-Circle array. Point "Front" goes to the L-network, and "Back" directly to the input of the feed system (see Fig 11-84).**

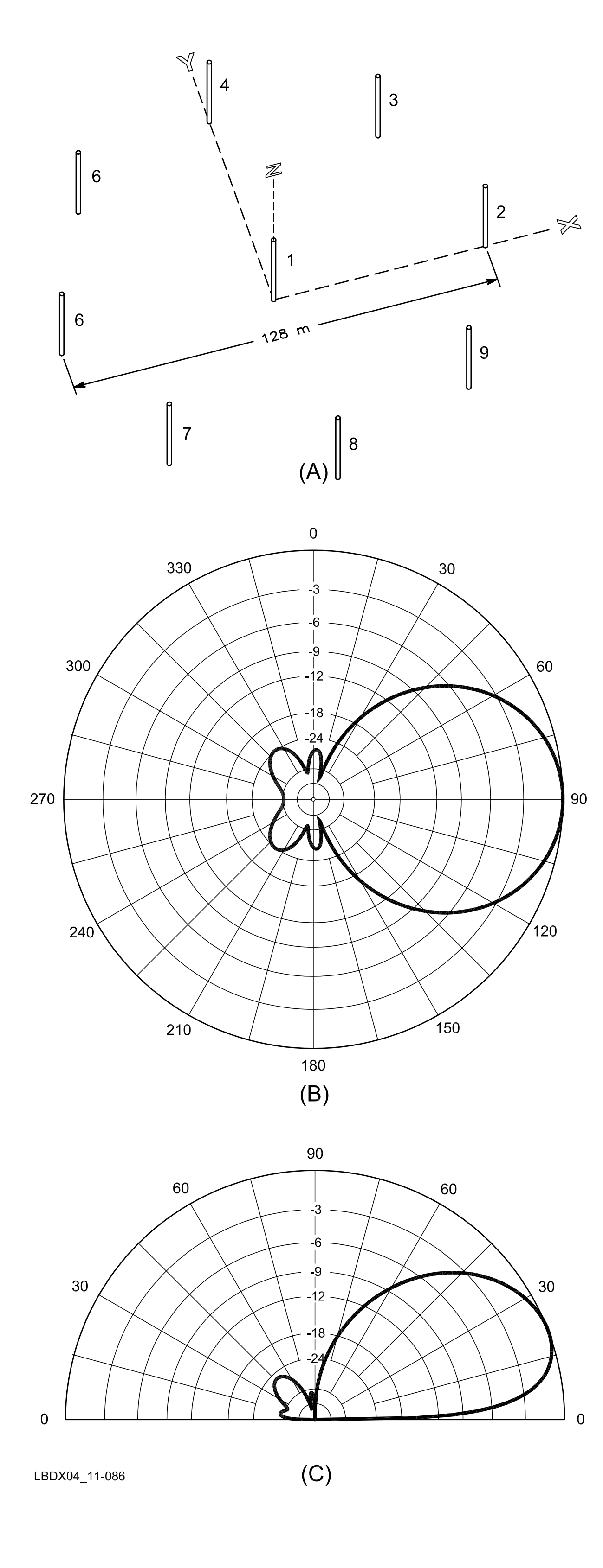



**Fig 11-86—Layout of the 9-Circle, along with the horizontal (at a 20° elevation angle) and vertical radiation patterns obtained with the larger 9-circle, which measures 128 meters in diameter.**

Feed impedances:
Z (Element 6) = −27.4 + *j* 2.2 Ω (back)
Z (Element 5) = Z (Element 7) = 12.3 – *j* 12 Ω (back, side-by-side)
Z (Element 4) = Z (Element 8) = 58.7 – *j* 26 Ω (middle, outer)
Z (Element 1) = 36.7 + *j* 3.1 Ω (middle, center)
Z (Element 3) = Z (Element 9) =75.9 – *j* 2.3 Ω (front, side-by-side)
Z (Element 2) = 112.5 + *j* 157 Ω (front, tip)

Note that the impedances include 2-Ω equivalent loss resistance in each element. I also modeled a smaller version of the 9-Circle, with a diameter of 80 meters vs 128 meters for the “big” one above. As expected the gain is down somewhat (by 0.7 dB), as well as the 3-dB forward beamwidth (69.6° vs 58.5°). This resulted in a somewhat lower RDF (12.39 vs. 12.99 dB). The behavior in the back is almost as spectacular as for its larger brother, resulting in a DMF of not less than 31 dB!

### 4.15.2. Data, Small 9-Circle array

Size = 80 meters diameter
Gain = 8.15 dBi
3-dB forward angle: 69.6°
RDF = 12.39 dB
DMF = 31.0 dB
Both RDF and DMF are nothing less than spectacular. Over very good ground (ε = 20 and conductivity = 30 mS), the gain is 9.9 dBi.
Feed currents:
Element 6 (back, tip): I = 1 /0° A
Element 5 and Element 7 (back, side-by-side): I = 1.4 /–100° A
Element 4 and Element 6 (middle, outer): I = 1 /–200° A
Element 1 (middle, center): I = 3 /–200° A
Element 3 and Element 9 (front, side-by-side): I = 1.4 /–300° A
Element 2 (front, tip): I = 1 /–40° A
Feed impedances:
Z (Element 6) = −18 – *j* 7.5 Ω (back)
Z (Element 5) = Z (Element 7) = 10.9 – *j* 20.1 Ω (back, side-by-side)
Z (Element 4) = Z (Element 8) = 54.9 – *j* 11.1 Ω (middle, outer)
Z (Element 1) = 27.6 + *j* 2.2 Ω (middle, center)
Z (Element 3) = Z (Element 9) = 54.2 + *j* 48.8 Ω (front, side-by-side)
Z (Element 2) = −77 + *j* 104.7 Ω (front, tip)

Calculations are done including an equivalent ground loss resistance of 2 Ω in each element.

### 4.15.3. The K9DX 9 Circle near Chicago

John, K9DX, built his 9-Circle using 27-meter long elements (Titanex 160HD). He used 120 quarter-wave long radials on each element and the tradeoff caused by these shorter element is nil. John tuned the element with a high-Q coil at the bottom of each element. See **Fig 11-87**. Of course, the feed impedances are different from those shown above.

K9DX supplied these impedances (including the loading coil, which has an inductive reactance of 120 Ω):

Z (Element 6) = −16.6 – *j* 1.0 Ω (back)

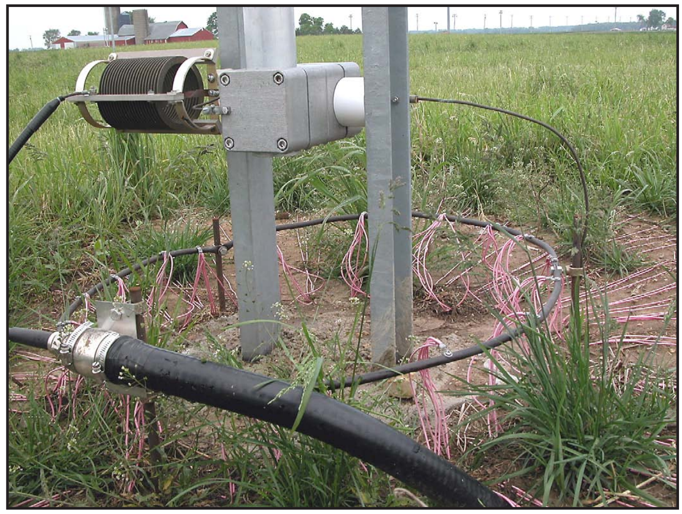


**Fig 11-87—Base of one of the elements of the K9DX 9-Circle. Note the high-Q loading coil and the ring (1-meter diameter) made of 10 mm copper, to which all of the 120 quarter-wave radials are connected. K9DX uses $1\frac{5}{8}$-inch coax for the $3\lambda/4$ feed lines.**

Z (Element 5) = Z (Element 7) = 5.9 – *j* 7 Ω (back, side-by-side)
Z (Element 4) = Z (Element 8) = 30.3 – *j* 15 Ω (middle, outer)
Z (Element 1) = 18.7 + *j* 0 Ω (middle, center)
Z (Element 3) = Z (Element 9) = 52.2 + *j* 48.8 Ω (front, side-by-side)
Z (Element 2) = 53.7 – *j* 127 Ω (front, tip)

### 4.15.4. The K9JA 9-Circle in Virginia

The 80-meter 9-Circle at K4JA uses full-sized quarter-wave elements, made of 5-cm OD aluminum tubing elements. **Fig 11-88** shows K4JA’s elegant array.

### 4.15.5. Feed system, 9-Circle array

Modeling an impressive array like a 9-Circle is one thing. Building it and making it to work like it does on the computer screen is a totally different thing! So far as designing a feed system, there are many roads that lead to Rome.

In the design shown in **Fig 11-89** I made use of two L-networks and three transformers (one 9:1 impedance ratio transformer and two 1:1 transformers). Note that this is not the feed system that K9DX is actually using.

John, K9DX uses a motor-driven rotary switch with heavy contacts on nine ceramic wafers. Each wafer is connected to one element. This system ensures minimum inductance, but is expensive if you need to buy the switch new. It is available from Multi-tech Industries, (**www.multi-tech-industries.com/**). See **Fig 11-91**. K4JA changes directions on his 80-meter 9-Circle using small vacuum relays mounted to an aluminum plate, as shown in **Fig 11-92**.

The feed system in Figs 11-89 and 11-90 uses only two L-networks and no additional coaxial cables to obtain the required phasing angles. The center element (with a relative feed current of 3) is fed directly from the input terminals through a 180° phase-reversal transformer. This is better than a 180°-long piece of coax because it is frequency-indepen-

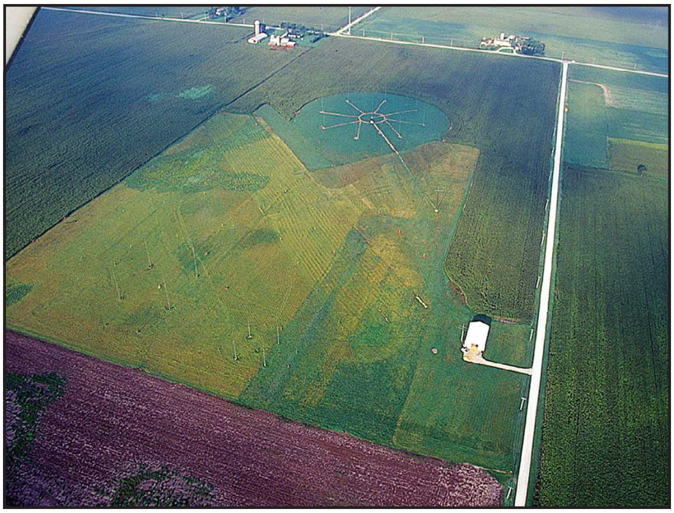


**Fig 11-88—Arial picture of the K9DX 9-Circle array. The circular terrain, including the radials that extend 40 m from the array itself, has a diameter of 208 m (two soccer fields long!). The brown stripes are the feed line trenches that had not grown over when the picture was taken.**

dent, and if well-made has little loss and a phase delay of a little bit less than 180° (see Section 4.15).

The two elements that are fed in phase with the center element, but with 1/3 of the current magnitude, are also fed via an identical 180° phase-inverter transformer, and through a 9:1 impedance ratio transformer (which is a 3:1 voltage ratio, and a 3:1turns ratio). This ensures that these elements get three times less feed current compared to the center element.

The front and the back elements are 360° out-of-phase, which means that they are in-phase. Since they have a feed current magnitude of 1, they can be connected directly to the output (low Z side) of the 9:1 transformer.

So far we have the three center elements fed at –180°, the front element at –360° and the back element at 0°. This is a very simple setup using only broad-band transformers.

We will now design two L-networks that take care of the proper phase angle and magnitude for feeding the remaining elements—two directors and two reflectors. These are all fed at a relative current magnitude of 1.66. For this we use the *Lahlum.xls* spreadsheet. The inputs of the L-networks are connected to the line that feeds the center element, which has a phase angle of –180°. The two directors require a phase angle of –270°, so will require a θ of –90°. The k factor is 1.66/3 = 0.55. The two reflectors require a phase angle of –90°, which means that for this L-network θ = –90 –(–180) = + 90° – 360° = –270°. The same k factor applies (0.55).

Fed this way, the array appears to have a total array feed impedance of 19 Ω, which is quite acceptable. Using 75-Ω cable (see Fig 11-90) for the current-forcing feed lines, we obtain a feed impedance of 40.4 + *j* 19.7 Ω. With just a paral-

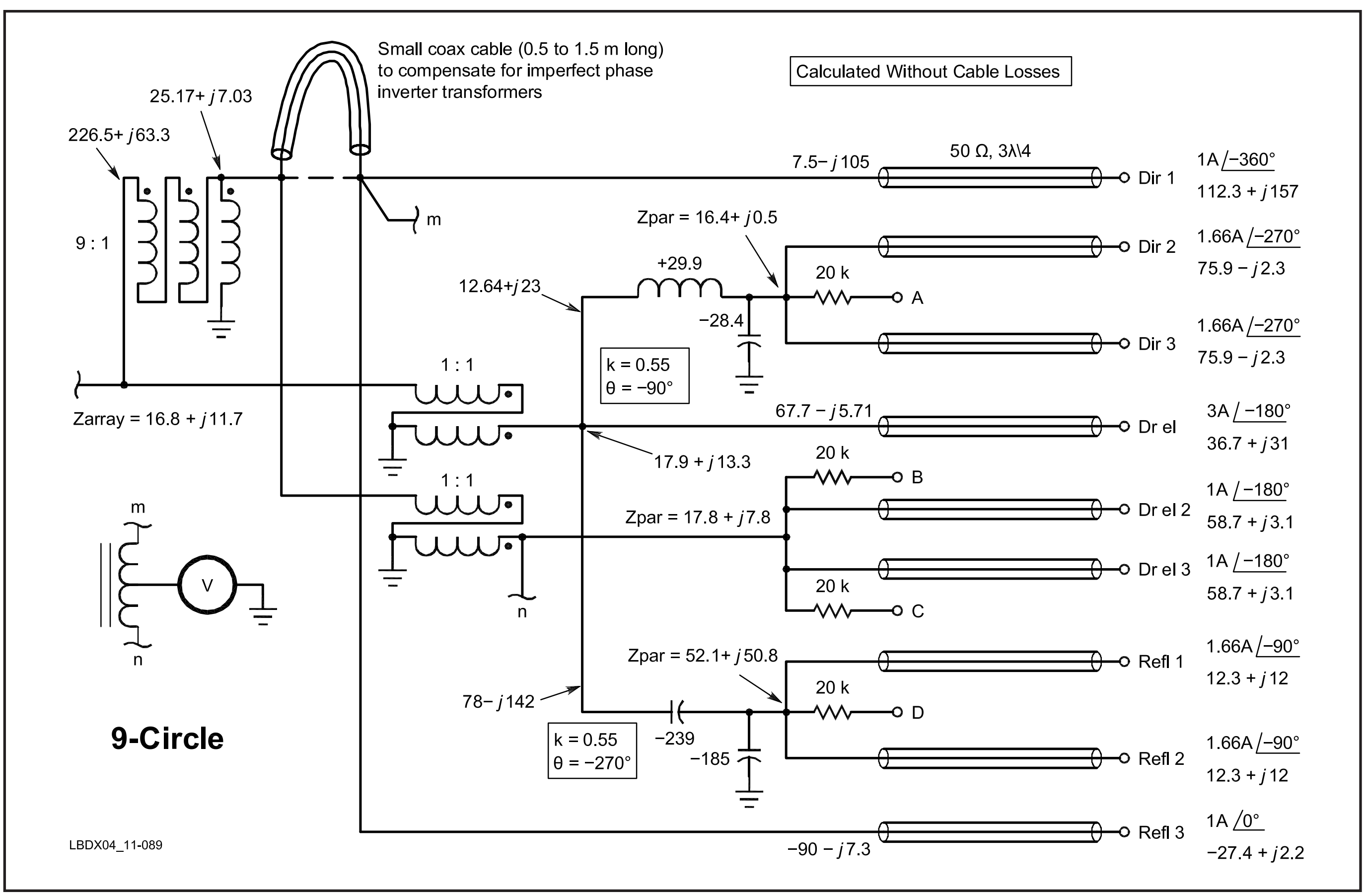



**Fig 11-89—Feed system for the quadrature-fed 9-Circle array based on 50-Ω feed lines. The calculations were done without taking cable losses into account. See text for details.**

lel capacitor (reactance = −102.6 Ω), the feed impedance turn into exactly 50 Ω (lucky strike!). With "real cables" the impedance will be a little higher, maybe 52 to 55 Ω, depending on cable losses.

It turns out that the values calculated for the 2 L-networks are very normal and quite manageable. To adjust the components to obtain the right phase angle and feed current magnitude, we can use the hybrid-coupler adjustment system developed by W1MK (see Section 3.6.2).

In Figs 11-89 and 11-90 we see four voltage divider resistors (20 kΩ) installed. All we need to do is connect a hybrid coupler between points A and B and another one

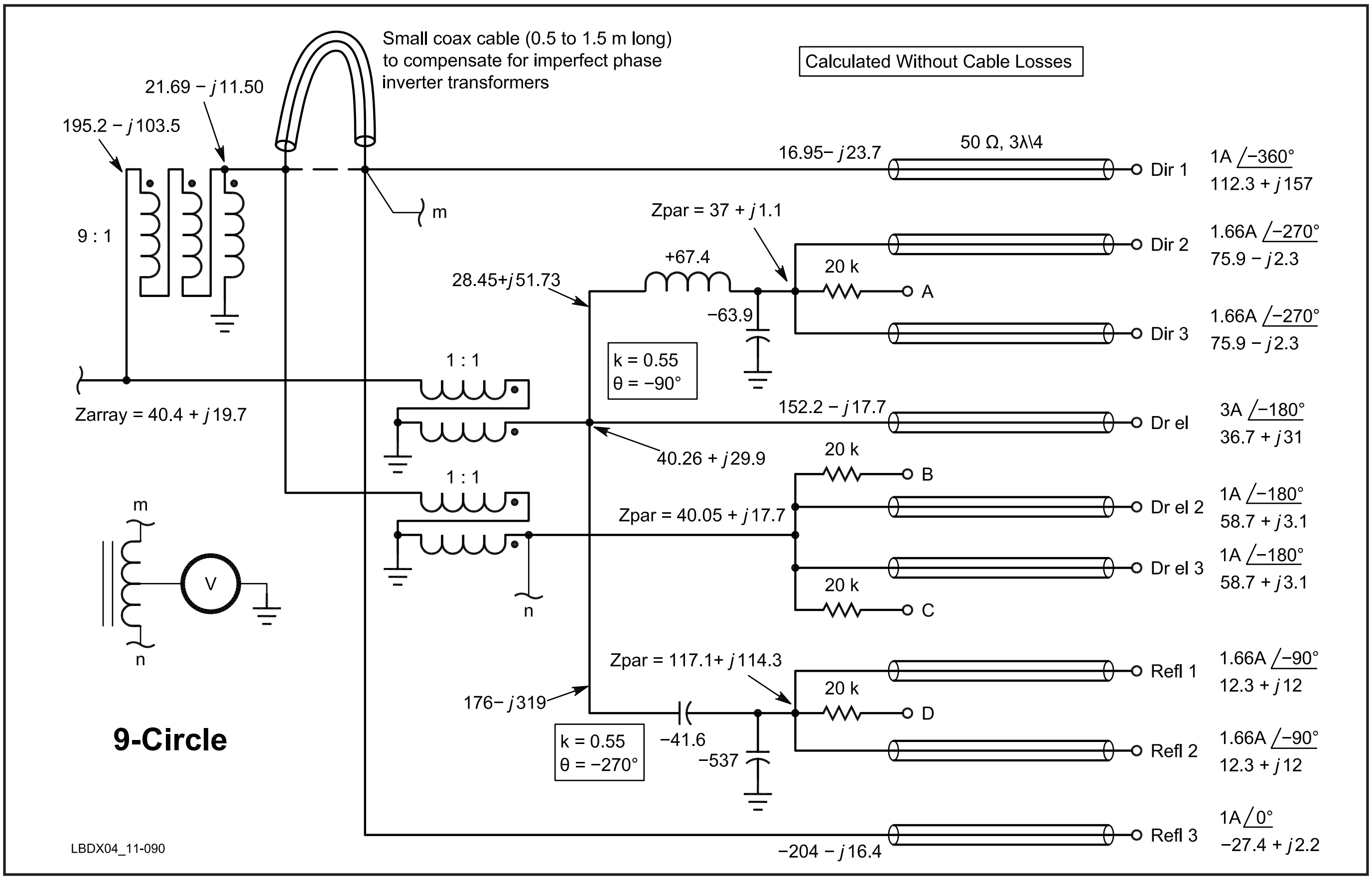



**Fig 11-90—Feed system for the quadrature-fed 9-Circle array based on 75-Ω feed lines. The calculations were done without taking cable losses into account. With real-world lines that have line losses the array feed impedance will be very close to 50 Ω.**

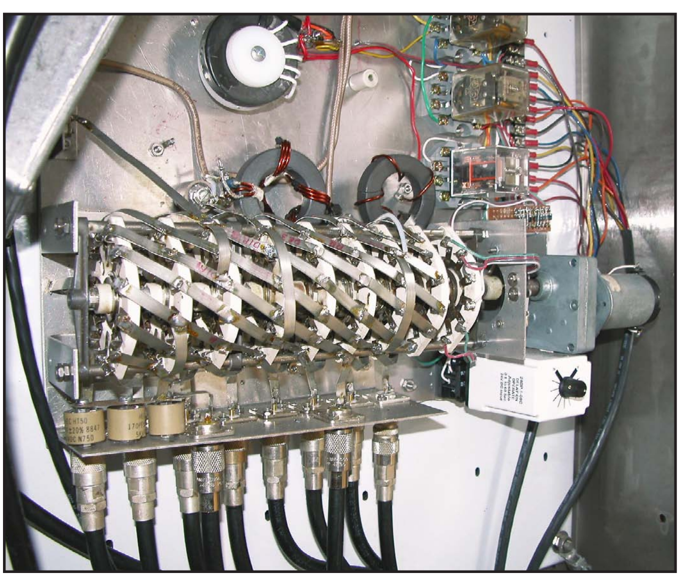


**Fig 11-91—A motor-driven rotary switch is used for direction switching at K9DX.**

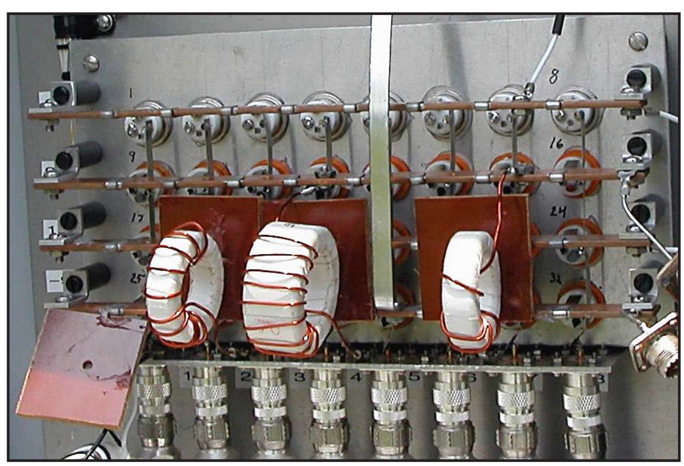


**Fig 11-92—K4JA uses a matrix of 8 by 4 small vacuum relays to do direction switching. This has inherently more inductance, but if you use the adjustable L-networks it should be possible to tune out the effect of the stray inductances in the wiring.**

between C and D. As the required phase angle difference is 90°, we need no extra lengths of coax to correct for non-quadrature phase angles. We will have to provide some attenuation in the legs going to the points B and C though. The voltage ratio is 3/1.66 = 1.81, or in dB: 20 × log 1.81 = 5.14 dB. We need two 50-Ω attenuators of 5.14 dB. A T-attenuator using 15-Ω resistors in the series branches and 82-Ω in the parallel branch will be very close.

If we use the detector/wattmeter designed by W1MK, we can adjust the values of the four components until we get a good null on the summed output. Bingo!

### 4.15.6. Transformer construction, 9-Circle array

K9DX, has gained a lot of first-hand experience building RF transformers required in this array. John winds most of his transformers with RG-303 single-shield Teflon coax. For the 9:1 unit he takes the shield from another piece of coax and slides it over the Teflon insulation to get three conductors. The core material used is 61 material, permeability of 125, 2.4-inch OD ($A_L$ = 171).

Not all of us have access to professional equipment to measure loss, impedance ratio and phase delay, so John developed a quick way of checking the transformer's performance. He terminates the transformer with the impedance it will normally see, then opens and shorts the output terminals. The input impedance with an open-circuit must go up by 10 times, and the impedance with a short-circuit must go down by 10 times. If it meets these requirements, it will likely measure OK on sophisticated test equipment!

### 4.15.7. 180° phase-shift transformers, 9-Circle array

K9DX also found that unless the transformer is allowed to be lossy (when wound with too few turns), the resulting delay will be 180° plus the delay inherent in the length of coax or wire involved. The usual result is in the183° to 185° range. **Fig 11-93** is a photograph of one of K9DX's 180° phase-inverting transformers. But there is a way to compensate for this imperfection.

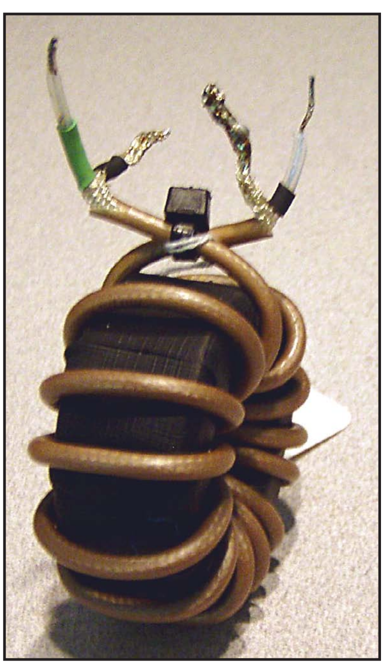


**Fig 11-93—K9DX made a 180° 1:1 phase-reversal transformer wound using 13 turns of RG-303 coax on a two side-by-side, 2-inch OD #61 cores.**

If we can introduce a similar amount of "extra phase" shift (3° to 5°) in the branches not fed via the 180° transformer(s) we can compensate for the imperfect phase-reversal transformers. A short piece of coax (0.5 to 1.5 meters long, the exact length will depend upon the SWR on the line) has this compensating effect. To adjust the short coax length to obtain 180° phase delay in the system, we can use a small push-pull (balanced) transformer. Connect the balanced inputs (m and n in Figs 11-89 and 11-90), and adjust the length for minimum voltage between the center tap and ground. You can use the detector-wattmeter described in Section 3.6 as an RF voltmeter.

**Fig 11-94—9:1 transformer made by John, K9DX. The transformer has four trifilar turns. The trifilar wiring is made by using a small Teflon coax equipped with a second shield. See text for details.**

K9DX commented: "Th*is 'over 180° problem' is one of the reasons I will probably stick to 180° coaxial lines in my 80 meter system. The lines can be adjusted to hit the delay right on the nose. Of course the downside is that their length must be changed between phone and CW, which adds more complexity*."

A 9:1 wide-band transformer built by K9DX is shown in **Fig 11-94**. **Fig 11-95** shows a wire-connection diagram for this transformer.

### 4.15.8. Bandwidth, 9-Circle array

A major issue in designing a feed network is bandwidth. It is relatively easy to adjust the phase and magnitude of the antenna currents at one frequency, but depending on the feed system used, things can fall apart rapidly when the frequency is changed.

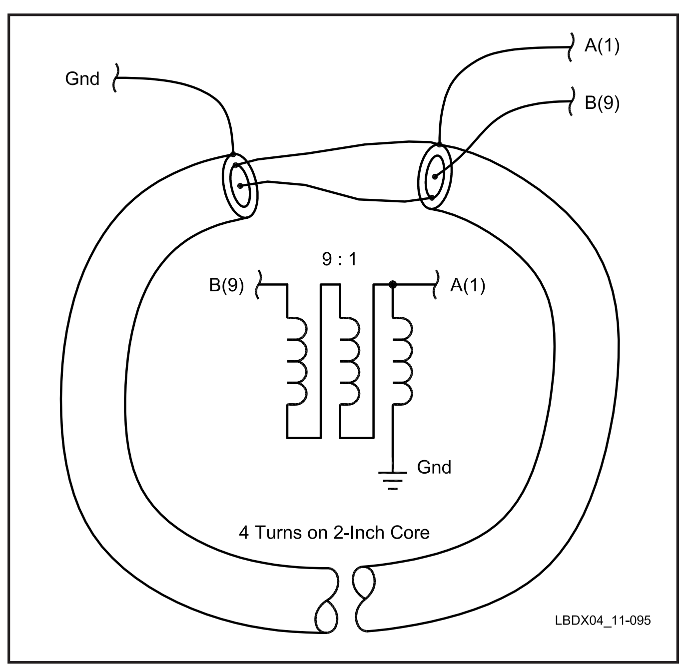



**Fig 11-95—Construction of 9:1 transformer. See text for details.**

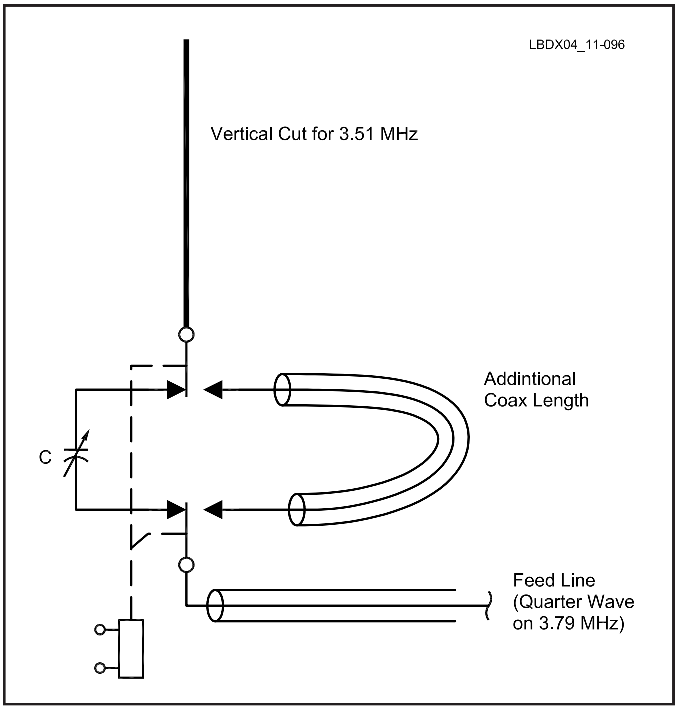



**Fig 11-96—To cover both the CW and the Phone ends on 80 meters, you must re-resonate the elements and change the length of the current-forcing feed lines.**

You can always make a feed system that does exactly what you engineered, at least on the frequency you engineered it for! If you set out to build an array like a 9-Circle, there is no room for compromises. Compromising to improved bandwidth is inevitably a losing battle. K9DX says that his 160-meter 9-Circle holds perfect directivity (down 30 to 40 dB off the side and in the back) over approx 30 kHz (1.8% bandwidth), which is adequate for that band.

On 80 meters, trying to cover the CW end and the Phone end (8% relative bandwidth) in one network is just *not* possible. The antenna elements become the wrong lengths, the phase shifting networks (coaxial or L-networks) move far away from their design center, and the current-forcing feeds destroy the amplitude and phase relationships. Note that is holds true not only for the 9-Circle but for all phased arrays that we want to operate on both the CW- and the phone end of the 80-meter band.

The only good solution on 80 meters is to resonate the elements on both ends of the band (for example, with a small loading coil on the CW end, or with a series capacitor to tune a CW element to become a phone-band element). See **Fig 11-96**. We then adjust the 3λ/4 (or λ/4 in case of "smaller" arrays) current-forcing lines to the exact lengths needed, by inserting extra cable lengths when operating at the CW end of the band.

With a quarter-wave current-forcing feed line, going from 3.5 to 3.8 MHz with a line cut for 3.8 MHz introduces

**Table 11-15**
**Summary, Phased Arrays**

| *Array Type* | *Gain (1) dBi* | *RDF dB* | *DMF dB* | *3-dB BW,Deg* | *Footprint (1)* | *Footprint (2)* | *Reference* |
|---|---|---|---|---|---|---|---|
| Single vertical | 0.9 | — | — | — | 0.16 | 0.4 | — |
| 2-ele end-fire quadrature | 4.2 | 8.1 | 12.3 | 177° | 0.32 | 1.0 | 4.1.1. |
| 2-ele end-fire optimized | 4.0 | 8.9 | 5.0 | 154° | 0.28 | 0.9 | 4.1.1 |
| 2-ele end-fire λ/8 spacing | 5.0 | 9.6 | 16.6 | 131° | 0.24 | 0.8 | 4.1.1. |
| 3-in-line quadrature | 5.3 | 9.2 | 17.9 | 143° | 0.50 | 1.3 | 4.4.1 |
| 3-in-line optimized | 5.6 | 11.2 | 27.8 | 98° | 0.40 | 1.1 | 4.4.1 |
| Triangle wide, quadrature | 5.0 | 8.8 | 14.3 | 150° | — | — | 4.6 |
| Triangle optimized | 5.8 | 9.8 | 16.9 | 129° | — | — | 4.6 |
| Four-Square, quadrature | 6.7 | 10.6 | 21.0 | 98° | 0.64 | 1.4 | 4.7.1 |
| Four-Square WA3FET | 7.3 | 11.4 | 24.4 | 85° | 0.64 | 1.4 | 4.8.2 |
| Four-Square, half dir. quadrature | 5.5 | 9.2 | 14.7 | 123° | 0.64 | 1.4 | 4.8.1 |
| Four-Square half dir. optimized | 5.9 | 9.7 | 16.5 | 113° | 0.64 | 1.4 | 4.8.2 |
| Broadside/end-fire, 0.55 λ spacing | 8.4 | 12.1 | 20.8 | 58° | 1.2 | 2.1 | 4.9 |
| 5-Rectangle | 5.9 | 9.8 | 19.4 | 122° | 1.2 | 2.1 | 4.10 |
| 5-Square | 7.7 | 11.9 | 23.0 | 78° | 0.8 | 1.7 | 4.11 |
| 5-Square, half direction | 6.2 | 10.5 | 18.5 | 102° | 0.8 | 1.7 | 4.11.2 |
| 6-Circle 60-m diameter | 7.6 | 11.7 | 25.5 | 78° | 1.0 | 2.0 | 4.12.2 |
| 6-Circle 80-m diameter | 7.8 | 11.6 | 26.3 | 75° | 1.5 | 2.6 | 4.12.1 |
| 8-Circle optimized | 9.2 | 13.3 | 21.8 | 58° | 2.4 | 3.8 | 4.14.2 |
| 9-Circle 128-m diameter | 9.1 | 13.0 | 31.7 | 70° | 2.8 | 4.3 | 4.15 |
| 9-Circle 80-m diameter | 8.15 | 12.4 | 31.0 | 70° | 1.5 | 2.5 | 4.15.2 |

Gain (1): in dBi over good ground (ε = 13, conductivity = 5 mS/m)
Gain (2): in dBi over Very Good Ground (ε = 30, conductivity = 30 mS/m)
Footprint (1): footprint in hectares with 20-meter long radials (1 hectare ~ 2.5 acres)
Footprint (2): footprint in hectares with 40-meter long radials

a phase shift error of no less than 7°, if the line is flat. With 3λ/4 current-forcing lines the phase error becomes three times that much on a flat line, or 21°. On a feed line with SWR (they all have SWR, sometimes very high such as on the "back" or the "front" elements of the array, the phase angle error for a 3λ/4 line cut for 3.5 MHz and operating on 3.8 MHz can be as much as 100° or 150°! This make it totally impossible to make the antenna work on both ends of the band.

### 4.15.9. Current-forcing feed lines, 9-Circle array

One solution for an 80-meter array is to resonate the elements on the CW end, and re-resonate them in the SSB band with a relay and a series capacitor. In addition you can add a short piece of coax to the quarter wave feed lines to make an exact λ/4-feed line also on 3.5 MHz. This is even more important if you are using 270°-long feed lines, where things move three times as fast off target as compared to using 90° feed lines.

### 4.15.10. Conclusion, 9-Circle array

The 9-Circle is a low-band array most of us can only dream of. It's an interesting subject where you let your imagination go and design your own feed system. There are numerous alternatives, and they all have pros and cons.

Building and owning a 9-Circle array is not for everyone: a 160-meter version requires over 4 hectares (~ 10 acres) of real estate! It also requires more-than-average knowledge in antenna matters and in electronics to design the feed system, to build it and to adjust it.

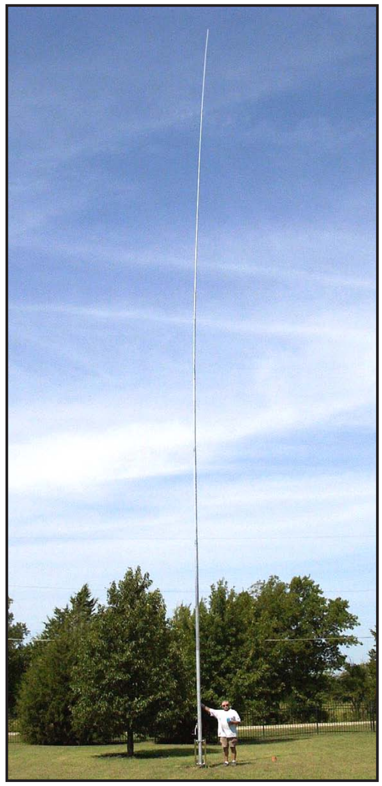


**Fig 11-97—Self supporting full-size 80-meter λ/4 elements are commercially available from WXØB at Array Solutions.**

## 4.16. Summary, Phased Arrays

**Table 11-15** gives a performance overview of the arrays covered in this chapter. Gain is given over "Average Ground." RDF and DMF are also listed. Interesting information is the footprint information, which is given for the array with 20-meter long radials, as well as for 40-meter long radials. The figures apply for a 160-meter antenna. For an 80-meter array, the footprints are four times smaller.

## 5. ELEMENT CONSTRUCTION

## 5.1. Mechanical Considerations

Self-supporting λ/4 elements are easy to construct on 40 meters. On 80 meters it becomes more of a challenge, but self-supporting elements are feasible even with tubular elements when using the correct materials and element taper. Tubular full-size λ/4 elements for 80 meters are available commercially from Arrays Solutions (**www.arraysolutions.com/**) as well as from Titanex (**www.titanex.de/**). See **Fig 11-97**. Lattice-type construction is commonly used, with tapering-diameter aluminum tubing at the top. On Topband, most quarter-wave vertical radiators are guyed towers. Rohn 25 tower is often used. See **Fig 11-98**. As it is highly recommended to series feed the elements of an array, the elements should be insulated from ground. This poses additional mechanical constraints on construction.

I used the ELEMENT STRESS ANALYSIS module of the YAGI DESIGN software (see Chapter 4 on software) to develop self-supporting elements for 40 and 80 meters that withstand high wind loads. As the element is vertical, there is no loading of the element by its own weight, which means that the same element in a vertical position will sustain a higher wind load than in a horizontal position. When using the ELEMENT STRESS ANALYSIS module, you can create this condition by entering a near-zero specific weight for the material used.

It will, however, be much easier if you plan to have at least one level where the vertical can be guyed. This will typically lower the material cost for constructing a wind-survival vertical by a factor of three or more. Finally, the element construction that is best for your project will be dictated to a large extent by material availability.

Needless to say, guying materials need to be electrically transparent guy wires (Kevlar, Phillystran, Nylon, Dacron, etc) or metallic guy wires broken up into small nonresonant lengths with egg-type insulators. Refer to *The ARRL Antenna Book* (Chapter 22), which covers this aspect in great detail.

**Fig 11-98—Roger Vermet, ON6WU, installing the base insulator on the Rohn 25 tower sections used for the 160-meter Four-Square at A61AJ in Dubai.**

All the array data in this chapter are for λ/4-full-size elements. Electrically longer elements are to be avoided at all times, since such elements would be no longer fed at a current maximum and design becomes more complicated. It is not crucial, however, to use full-size elements. Top-loaded elements that are physically 2/3 full-size length can be used without much compromise. Make sure, however, that all elements in the array use the same amount of top-loading. If not, special precautions have to be taken (see Section 6.5). If guyed aluminum-tubing elements are used, the top set of guy wires can be used to load the element (see Chapter 9 on vertical antennas). If the array must cover 3.8 MHz as well as 3.5 MHz, a small inductance can be inserted at the base of each vertical (make sure the loading coils are identical!) to establish resonance for all elements at 3.5 MHz.

The main cause of failure with guyed aluminum tubing elements is *buckling*. This usually happens when four conditions are met:

1. Distance between guying points is too long.
2. Thin-wall flimsy aluminum material (easy bending).
3. Too much vertical load on the mast (too much guy pulling).
4. Too much wind (bending in between guying points, eventually turning into buckling).

## 5.2. Shunt Versus Series Feeding

Shunt feeding the elements of an all-fed array is to be avoided in just about all cases. The matching system (gamma match, omega match, slant-wire match, etc) introduces additional phase shifts that are difficult to control. Such phase shifts will mess up the correct feed current in the antenna elements.

Only with arrays where all the feed impedances are identical can shunt feeding be applied successfully. The feed impedances of all elements of an array will be identical only when all the elements are fed in-phase (or 180° out-of-phase). Shunt feeding may be considered for such arrays if the vertical elements as well as the matching systems are identical (including the values of any capacitors or inductors used in the matching system).

If you feel tempted to use your tower loaded with HF antennas as an element of an array, be aware that you might be trying to achieve the impossible.

- The loaded tower may be electrically quite long, which could very well be a hindrance to achieve the required directivity (see Section 2).
- You will be forced to use shunt feeding, which is just about uncontrollable, especially if all elements are not strictly identical (which will hardly ever be the case with such loaded towers).

Loaded towers are just great for single verticals, but are more than a hassle in driven arrays.

## 5.3. Loaded Elements

It is not always possible to use full size λ/4 elements, and provided you install a very low-loss ground system, full-size elements are not really required. On 160 meter many arrays have been built with inverted L elements, although T-loaded elements will produce much better directivity.

### 5.3.1. Inverted-L elements

The inverted-L vertical is described in Chapter 9, Section 7. For a single vertical, where we really do not expect much directivity at all, the inverted-L is a good antenna. In an array however, where you really are mostly after gain *and* directivity, the horizontally polarized high-angle component radiating from the flat-top section of the inverted-L is a problem. T-loading, with horizontal or even sloping top-loading wires arranged symmetrically, will reduce the high-angle radiation from these loading wires to almost zero.

### 5.3.2. T-Loaded Vertical Elements

If the central tower is not high enough to support full-size λ/4 verticals from the sloping support wires, these verticals can be top-loaded with a sloping top-wire. The top-loading wires can be part of the support system, as shown in **Fig 11-99**. The vertical elements are loaded with sloping top-wires to resonate them at 3.8 MHz. The sloping support wires have the property of not producing any horizontally polarized signal, provided the lengths on both sides of the vertical are the same.

As long as the vertical wire is not shorter than about λ/8, the loaded verticals will produce the same results as full-size verticals, with only some reduction in bandwidth. Section 7.5 describes a 160-meter Four-Square with vertical elements that are not longer than 18.5 meters.

## 6. ON4UN 4-SQUARE ARRAY WITH WIRE ELEMENTS

## 6.1. The Mechanical Concept

An 80-meter 4-square takes a lot of room to put up, not to speak about one on 160 meters! I have installed a somewhat special version of the 80-meter Four-Square around my full-size λ/4 160-meter vertical. This design has become very popular since it was first published. From the top of the vertical I run four 6-mm (1/4-inch) nylon ropes in 90° increments, to distant support poles. These nylon ropes serve as support cables from which I suspend the four verticals. A single radial is directed away from the center of the square

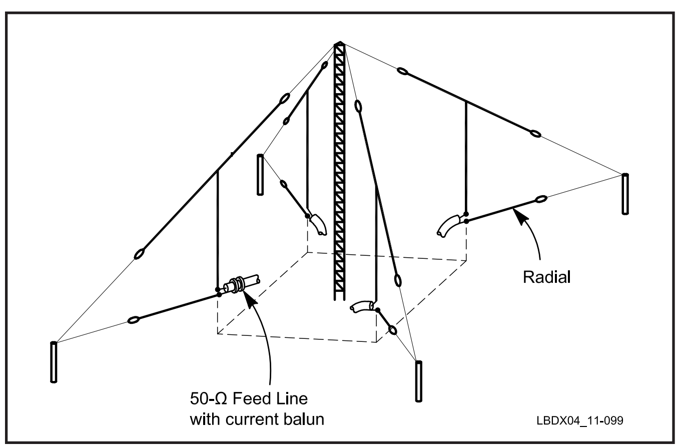



**Fig 11-99—If enough space is available, you can run four catenary cables from the top of the support tower to four supports mounted in a square. These catenaries support the verticals and their loading structures, if any. Sloping top-loading wires as shown here exhibit no horizontal radiation component, provided the length is the same on both sides of the vertical.**

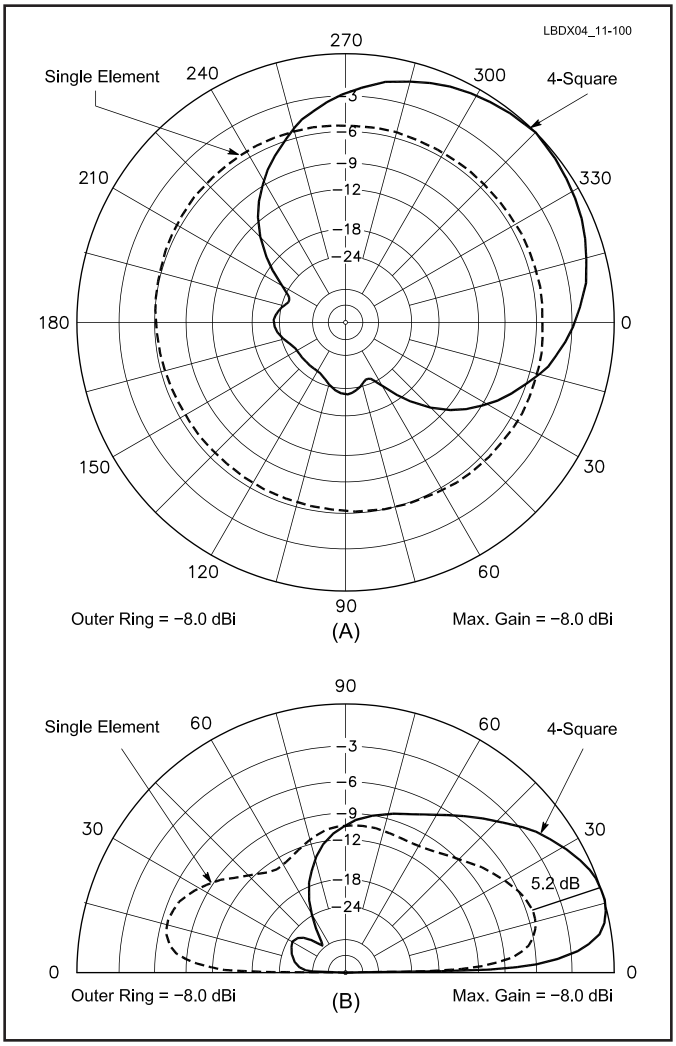



**Fig 11-100—Horizontal and vertical radiation patterns of the ON4UN Four-Square array with one elevated radial. Also shown is the pattern of a single vertical element. Both are modeled with a single radial per element, but over an extensive buried-radial system, 5 meters (17 feet) below the radial over very good ground. The buried radials are installed like spokes from the center of the square.**

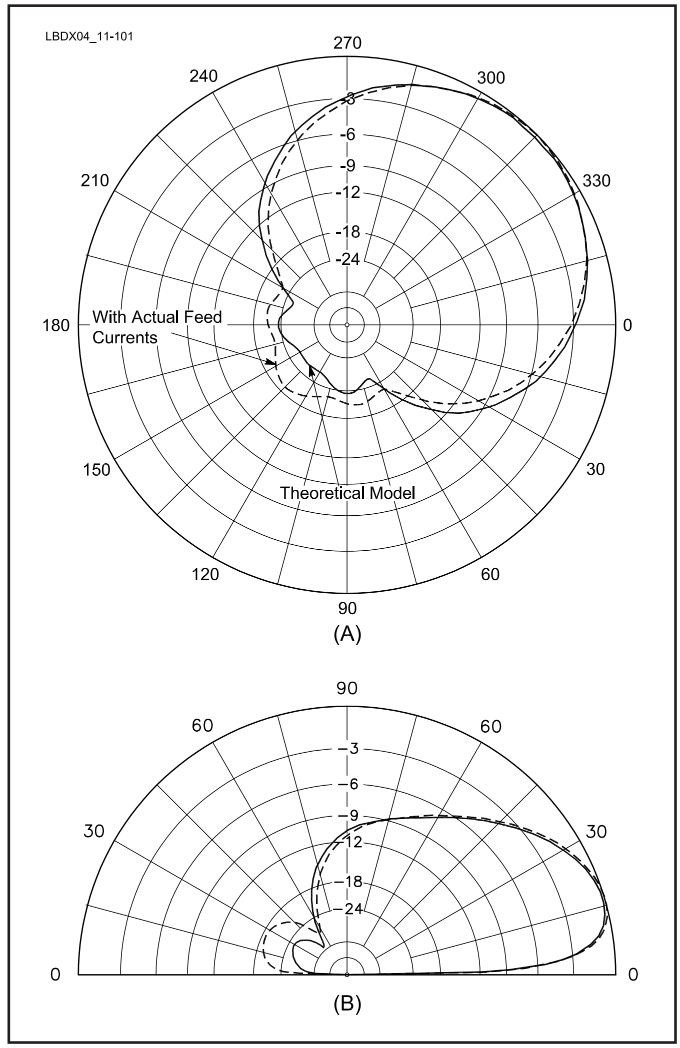



**Fig 11-101—Vertical and horizontal patterns (at a 20° elevation angle) for the theoretical currents and the actual currents at each element of the Four-Square array (see Fig 11-100). Note that although there are some significant current deviations (phase angle and magnitude) from the theoretical values, the array suffers only very slightly from these differences. These patterns were calculated using *EZNEC*.**

(where the 160-meter vertical is located). See **Fig 11-100**.

Since my support tower was high enough, I could manage full-size 80-meter vertical elements, with the feed point and the single radial 5 meters above ground. In this setup the single radial serves three purposes:

- It provides the necessary low-impedance connection for the feed line outer shield.
- It helps to establish the resonance of the antenna (which is not the case with a large number of radials or buried radials, where the resonance is mainly determined by the length of the vertical member).
- It provides some high-angle radiation. We can debate whether or not this is wanted, but in my particular case I wanted a fair amount of high-angle radiation as well, in order to be able to use the array successfully in contests, where shorter range contacts are also needed.

Just a single radial, without an extra ground screen on or in the ground will make you lose up to 6 dB of maximum achievable gain, depending on the quality of the ground below. I strongly advocate the use of extra radials or a ground screen under the verticals (see Chapter 9). In my particular case, there are some 250 radials (20 to 60 meters long) under my array, basically serving as the radial system for the 160-meter vertical that supports the array. With this extensive radial system, my 80-meter array exhibits a low-angle F/B of 20 to 25 dB, and still a very reasonable amount of directivity at relatively high angles (20 dB F/B at 60°).

The gain of this array at low angles is only 0.4 dB less than the gain of a Four-Square over a system using a perfect ground system. The slight drop is due to the power radiated at higher angles. The gain is 5.1 dB over a single element, which is very substantial.

**Fig 11-101** shows the horizontal and vertical radiation patterns for my array, as well as that of a single element over identical good ground (very good ground with 250 radials). Both the single vertical and the Four-Square use a single elevated radial for each vertical in the model.

The bottom ends of the four vertical wires are supported by steel masts that are located on the corners of a square measuring 20 meters, with the 39-meter tall 160-meter vertical right in the center of the square. The masts can be folded over for easy access to each element's feed point. The 80-meter vertical elements are 19.5 meters long. Together with a radial of 18.7 meters, the elements are resonant at 3.75 MHz. The individual elements of the array were measured to have a feed-point resistance of 40 Ω at resonance (3.75 MHz). I measured the impedance over a frequency range going from 2 to 5 MHz using an HP network analyzer with a Smith Chart display. Mutual coupling to other antennas and surrounding structures shows up on the Smith Chart as a kink or a dip in the impedance chart of one or more elements at a specific frequency. It is important that the impedance curves be as near alike as possible over the frequency range of interest, if the impedance variations when switching antenna directions are to be kept at a minimum. Section 3.3 deals with the problem of eliminating unwanted mutual coupling.

A word of caution: if the central supporting tower is a base-insulated tower, tuned to 160 meters, make sure that the tower is effectively grounded when you use the 80-meter Four-Square. If it is left floating, the central element can act as a half-wave element on 80 meters and can interfere heavily with the array. Grounding the central tower can be done in several ways, as discussed in Chapter 7 on special receiving antennas. The grounded 160-meter resonant tower in the center does not influence the performance of the Four-Square.

**Fig 11-102—A 10 × 10 × 3 cm plastic box is mounted on the top of the 5-meter support for each of the elevated wire verticals. Inside the box, the vertical wire and the single radial are connected to the feed line, which is equipped with a stack of 50 ferrite cores to remove any RF from flowing on the outside of the feed line. The box also houses the relay that switches the stub in and out of the circuit to lower the operating frequency to 3.5 MHz. The stub can be seen running along the steel support mast.**

## 6.2. Loading the Elements for CW Operation

To make the antenna cover the CW end of the band as well, I use a stub (or linear loading section), inserted in the radial at the feed point, to shift the resonance of the elements to 3.5 MHz. A small box is mounted on top of each mast. All connections (to the vertical element, radial, and feed lines) are made inside this box. The box also contains a vacuum relay that can switch the stub in and out of the circuit. The stub is supported by stand-off insulators along the metal support mast (see **Fig 11-102**).

The calculated reactance of the stub is +130 Ω. Using 3-mm-OD (AWG #9) copper wire with a spacing of 20 cm, the length of the stub turned out to be 2.25-meters long to lower the resonant frequency to 3.505 MHz. The same stub, when shortened to 75 cm resonates the element at 3.65 MHz. A nice feature is that the resonant frequency can be changed anywhere between 3.5 and 3.75 MHz by using a movable shorting bar across the stub. This way, you can create different operating windows on 80 meters. A relay can be used to switch the 3.65-MHz shorting bar in and out of the circuit, making the window selection remotely controlled. The direction control box contains a three-position lever switch, which selects the three band segments.

## 6.3. The λ/4 Feed Lines

Each element is fed via an electrical λ/4 of coaxial feed line with a current balun (50 stacked ferrite beads on a short length of small-diameter Teflon coax) at the feed point. The feed lines were cut to be λ/4 at 3.75 MHz. If a perfect 90° phase shift is desired at 3.5 MHz, the feed lines can be lengthened by a 1-meter long piece of coax (VF = 66%).

## 6.4. Wasted Power

In the case of a quadrature-fed Four-Square with 90° side dimensions, the use of 50-Ω feed lines results in combined feed-line impedances that load the ports of a 50-Ω hybrid coupler heavily (22 + *j* 19 Ω and 23 + *j* 17 Ω). This results in up to 11% of the power being dissipated in the dummy load resistor. With 75-Ω feed lines, these impedances are much higher (51 + *j* 37 Ω and 48 + *j* 43 Ω), which results in much less power being dissipated in the load resistor (4%). Again, the main parameter that determines the operational bandwidth of an array fed with a hybrid coupler is the amount of power being dissipated in the load resistor.

## 6.5. Gain and Directivity

Section 3.6.2 explains that the feed current in the elements can be assessed by measuring the voltage at the end of the quarter-wave feed lines going to the elements. When I initially built the array, I used a vector voltmeter to measure the voltages.

With the current-forcing method employed, the relative element feed-current requirement for equal-magnitude in quadrature phase is voltages with equal magnitudes at the ends of the λ/4 feed lines. (Here, $E = Z_0 \times I$ or $E = 50 \times V$ for a 50-Ω line). There was deviation from the theoretical values. As expected, the voltage magnitudes and phase angles were not exactly perfect. The voltage magnitude varied as much as 1.7 dB (41 V versus 50 V theoretically), while the phase angle was up to 13° off from the theoretical value for the 50-Ω feed-line case. In a pleasant surprise, even the relatively important deviations for the 50-Ω impedance case influenced the directivity pattern and gain only very marginally.

Using 75-Ω, λ/4 (or 3λ/4) feed lines does not make this a 75-Ω system. In this particular case I am still using a hybrid coupler with a 50-Ω nominal design impedance. The 75-Ω cables are used only because they transform the element feed-point impedances to more suitable values, resulting in less power dissipation in the dummy resistor.

## 6.6. Construction

In the original ON4UN layout, the Comtek Systems 50-Ω hybrid coupler (including a 180° phase-shift transformer) and the hybrid-coupler load resistor are located in a cabinet mounted at the base of the 160-meter vertical, in the center of the 80-meter Four-Square. The Comtek unit was removed from its normal housing and the PL-259 hardware was replaced with N connectors. The cabinet also contains the relay that switches the feed line between the 160-meter vertical and the 80-meter Four-Square. In 2004 I modified the cabinet to include a Lahlum/Lewallen system manufactured by Array Solutions. Both can be instantly selected from the shack. See **Fig 11-103.**

In order to know at all times how much power is being dissipated in the dummy load, I added a small RF detector to the dummy-load resistor and fed the dc voltage into the shack, where the relative power is displayed on a small moving-coil meter mounted on the homemade direction-switching box. The box also contains the switch to select the subbands, mentioned above. In addition, a level-detector circuit is included, using an LM-339 voltage comparator, which turns on a red LED if the dissipated power goes above a preset value. **Fig 11-104** shows the schematic of the system and **Fig 11-105** shows the actual switch box.

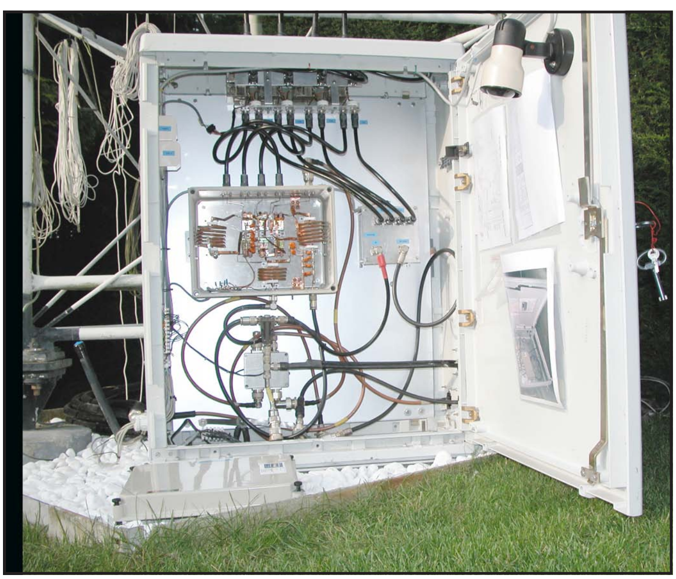


**Fig 11-103—Cabinet located at the base of the 160-meter vertical, housing the Comtek hybrid-coupler feed system and directional switching circuitry for the Four-Square array, along with a Lahlum/Lewallen feed system built by Array Solutions. The feed lines to the elements are 75-Ω 1/2-inch cable (Vf = 0.88) and cut for the system to be λ/4 on 3.775 MHz. In the cabinet a short length (0.97 meters) of RG-11 is added to make the feed lines exactly λ/4 on 3.51 MHz. The 180° delay line for the Array Solutions box is made of 7/8-inch 50-Ω hard line. It too is switched to be an exact half wave on 3.51 MHz or 3.885 MHz with a 3-meter section of 1-inch hard line (coiled up under the relay box).**

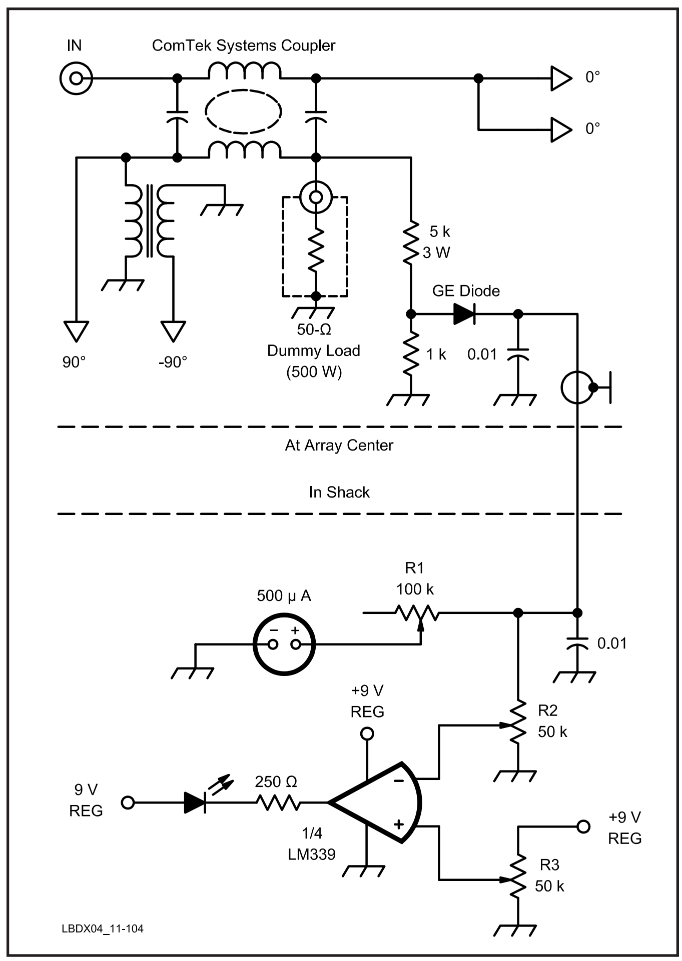



**Fig 11-104—Schematic diagram of the RF detector and voltage comparator used to monitor the RF into the hybrid terminating resistor. The LED will switch on if the voltage coming from the detector is higher than the preset voltage supplied by the potentiometer R3. R1 adjusts the sensitivity of the indicator, and the R2 sets the alarm level.**


**Fig 11-105—Array-direction switch box at ON4UN, with 1 direction switch for 40; 2 for 80. A switch selects the Comtek or Lahlum/WXØB system. Relative power and alarm circuit (Fig 11-104) is for Comtek. A lever switch selects one of three band segments on 80.**

## 6.7. Using the Lewallen Phasing Feed System

As explained earlier, the Lahlum/Lewallen system has two design advantages: There is no power lost in a dummy resistor and you can design the system for any phase angle and feed current magnitude. Non-quadrature feeding (eg, the configuration discussed in Sections 4.7.2 and 6.8.4) can give 1.0 dB of extra gain. Considering the power loss in the dummy load, 1.5 to 2 dB can be gained by going from a quadrature hybrid system to a Lahlum/Lewallen system.

In 2004 I rebuilt the 80-meter Four Square to use either feed system, switchable from the shack. When you consider a Lahlum/Lewallen system, please remember that a hybrid coupler system is a "plug-and-play" system, while a Lahlum/Lewallen system must be tuned and requires some test equipment.

## 6.8. Array Performance: Comparing the Feed Systems

### 6.8.1 Using the hybrid coupler

Judging the array performance by measuring its SWR is totally meaningless (see Section 3.4). With a hybrid coupler, my old array showed an SWR of less than 1.3:1 over the entire 80-meter band, wherever the resonances of the elements were.

Based on a wasted-power bandwidth criterion, however, the bandwidth of this array is 100 kHz without tuning for individual band segments. Fig 11-55 showed the wasted-power curve for my original Four-Square array with a single elevated radial at each element. The steepness of the dissipated-power curve is determined by the Q factor of the array elements. For my Four-Square, the elements are made of wire, so the Q is high and the bandwidth is narrow.

Also the fact that I use a single radial instead of a comprehensive buried radial system adds to the sharpness of the curve. While changing the frequency away from the design frequency, the single radial (just like the vertical element) will introduce reactance into the feed-point impedance, which would not be the case with a buried radial system.

Practically speaking, this array was by far the best antenna I had ever had on 80 meters. On-the-air tests continuously indicated that the signal strength on DX was ranked with the best signals from the European continent. As far as directivity is concerned, it is clear that the array had a nice wide forward lobe, and that the relative loss half-way between two adjacent forward lobes was hardly noticeable (typically 2 dB). Long-haul DX very often reported, "You are S9 on the front and not copyable off the back." Even on high-angle European signals there was always a good deal of directivity with this array (typically 15 dB).

However, it was not a good receiving antenna when compared to the range of Beverages I have at my QTH. One of the reasons, of course, is the single elevated radials, which causes a big bulge in the vertical radiation pattern at high angles. As explained before, this was done on purpose, so that the antenna would radiated a reasonably strong signal at high angle as well, which is a real asset when working contests.

### 6.8.2. ON4UN Four-Square array, data for quadrature feed

Design frequency: 3.7 MHz
Length of verticals: 18.7 meters (2-mm OD wire)
Length of radials: 21.2 meters
Height of feed point/radial: about 5 meters
Feed currents:
I1 = 1 /– 180° A
I2 = I3 = 1 /-90° A
I4 = 1 /0° A
Gain: 5.2 dBi (over good ground)
3-dB beamwidth: 102°
F/B: 17 dB
RFD (receiving directivity factor): 9.08 dB
DMF: 13.3 dB
The calculated feed-point impedances are:
Z1 = 52.5 + $j$ 52 Ω
Z2 = Z3 = 34 + $j$ 0 Ω
Z4 = 7.5 – $j$ 2.5 Ω

### 6.8.3 Using a Lahlum/Lewallen feed system in a quadrature feed.

**Fig 11-106** shows the Lahlum/Lewallen network for the 80-meter array at ON4UN, when the Four-Square is fed in quadrature. These calculations were done without cable losses since the differences are minute.

### 6.8.4. Using a Lahlum/Lewallen feed system in a non-quadrature feed

The Lahlum/Lewallen feed system enables us to feed the elements with whatever feed currents and phase angles we want. A half hour using *EZNEC* resulted in a design that uses the same geometrical layout of elements but that represents a worthwhile improvement over the equal-current/quadrature-phasing configuration.

#### *6.8.4.1. Array data*

Design frequency: 3.7 MHz
Length of verticals: 18.7 meters (2-mm-OD wire)
Length of radials: 21.2 meters
Height of feed point/radial: about 5 meters
Feed currents: I1 = 0.7 /0° A
I2 = I3 = 1 /–110° A
I4 = 1.4 /–220° A
Gain: 6.2 dBi—This is 1 dB better than the same array fed in quadrature!
3-dB beamwidth: 88°

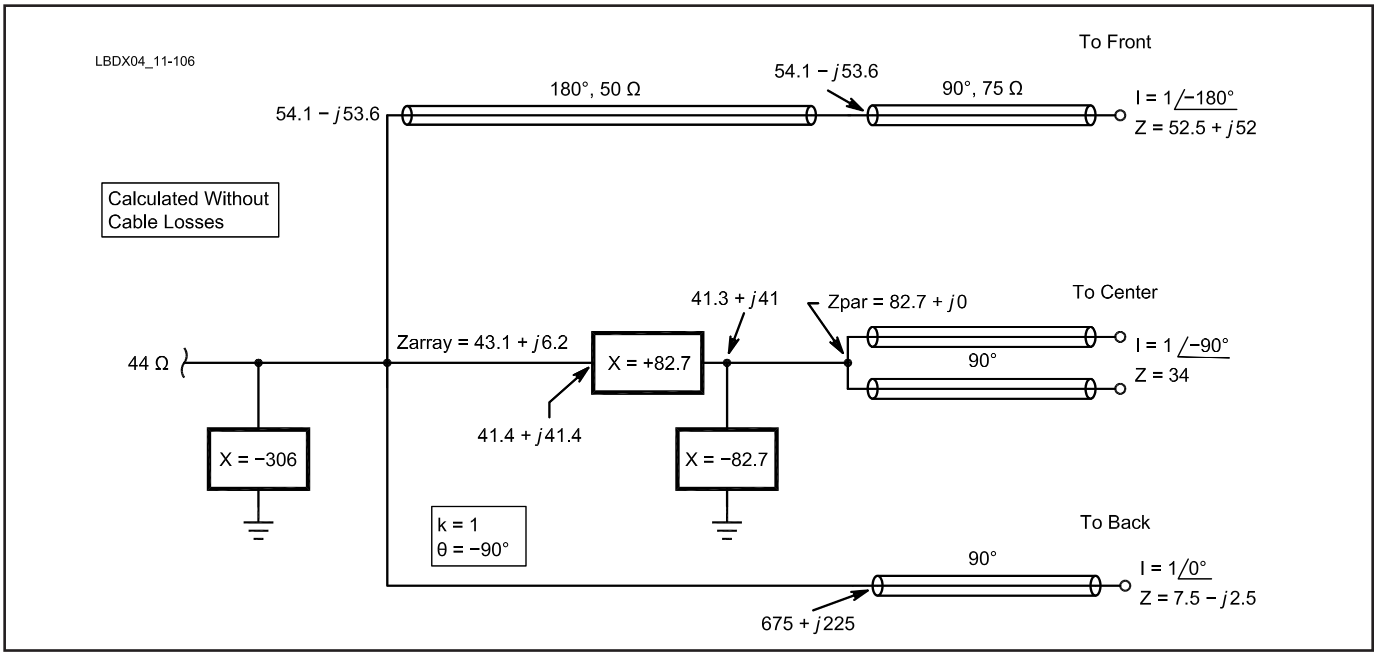



**Fig 11-106—The 4-element wire Four-Square at ON4UN was modeled in *EZNEC* and the impedance data used to design a quadrature-fed Lahlum/Lewallen feed system. The front element is fed via a 180° long phasing line, which could be replaced by a 1:1 phase-inverting transformer, as is done in the Comtek hybrid coupler.**

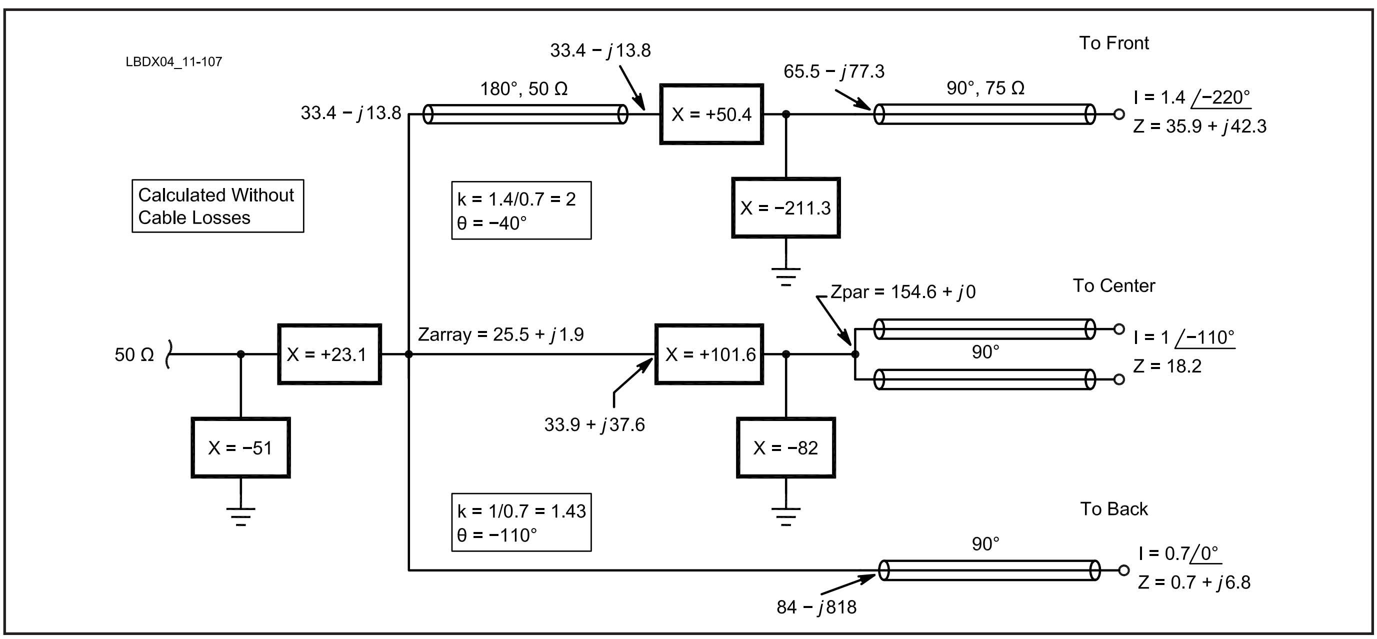



**Fig 11-107—Optimized Lahlum/Lewallen feed system for the ON4UN Four-Square.**

F/B: 21 dB
RFD: 9.57 dB
DMF: 14.4 dB
The calculated feed-point impedances are:
$Z1 = 0.7 + j\,6.8\ \Omega$
$Z2 = Z3 = 18.2 + j\,0\ \Omega$
$Z4 = 55.9 + j\,42.3\ \Omega$

**Fig 11-107** shows the new Lahlum/Lewallen feed system designed for this array. The values of the two LC network components can be adjusted using the test setup described in Section 3.6.2.

### 6.8.5. T-Loaded vertical elements

If the central tower is not high enough to support full-size quarter-wave verticals from the sloping support wires, these verticals can be top-loaded by a sloping top-wire. The top-loading wires can be part of the support system, as shown in Fig 11-99. The vertical elements are loaded with sloping top-wires to resonate at 3.8 MHz. The sloping-support wires have the property of not producing any horizontally polarized signal, provided the lengths on both sides of the vertical are the same.

As long as the vertical wire is not shorter than 2/3 full size (approximately 15 meters), the loaded verticals will produce the same results as the full-size verticals, with only a small reduction in bandwidth and gain.

## 7. ARRAYS OF SLOPING VERTICALS

In the chapter on dipoles, I describe a vertical half-wave dipole, as well as a sloping half-wave dipole and its evolution into a quarter-wave vertical with one radial. Sloping verticals are well suited for making a Four-Square array from using a single, tall tower as a support. In all these arrays the elements should be arranged in such a way that the feed points are located on a square measuring λ/4 on the side.

### 7.1. Four-Square Array with Sloping λ/2 Dipoles

A Four-Square array made of four λ/2 dipole slopers requires a support that is about 0.36 λ tall, 55 meters high at 1.84 MHz. The four feed points make a square measuring λ/4 on each side. **Fig 11-108** shows the layout for a 160-meter system. Fig 11-108B shows a side view of a single dipole, which includes the droop due to weight of the feed line.

### 7.2. Array data, Four-Square with Sloping λ/2 Dipoles

Quadrature feeding does not result in a very good pattern. Through modeling using *EZNEC* I came up with the following design for 1.8 MHz:

Feed currents: I1 = 1 /0° A
I2 = I3 = 1 /–139° A
I4 = 1 /–263° A

Gain: 4.56 dBi over good ground (about 4 dB over a single vertical)
3-dB beamwidth: 87° (vs 108° if quadrature fed!)
RDF (receiving directivity factor): 8.06 dB
DMF: 11.9 dB
The feed-point impedances calculated including a grounded

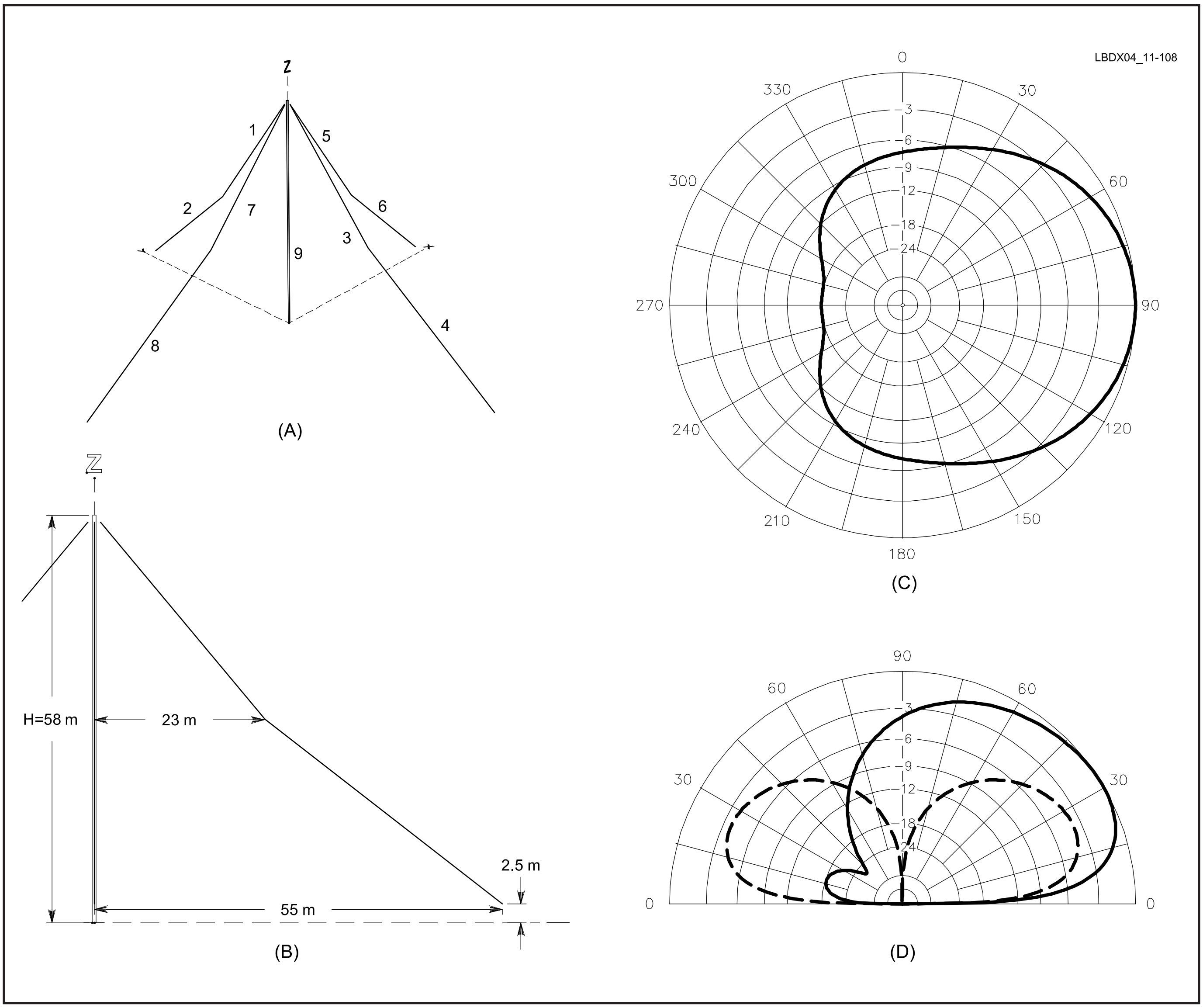



**Fig 11-108—At A, layout of 80-meter Four-Square using half-wave dipoles suspended from a single tall tower. At B, layout of one of the sloping dipoles, showing the droop due to the weight of the feed line. At C and D, horizontal and vertical radiation patterns for this array. The vertical pattern is compared with that for a single vertical.**

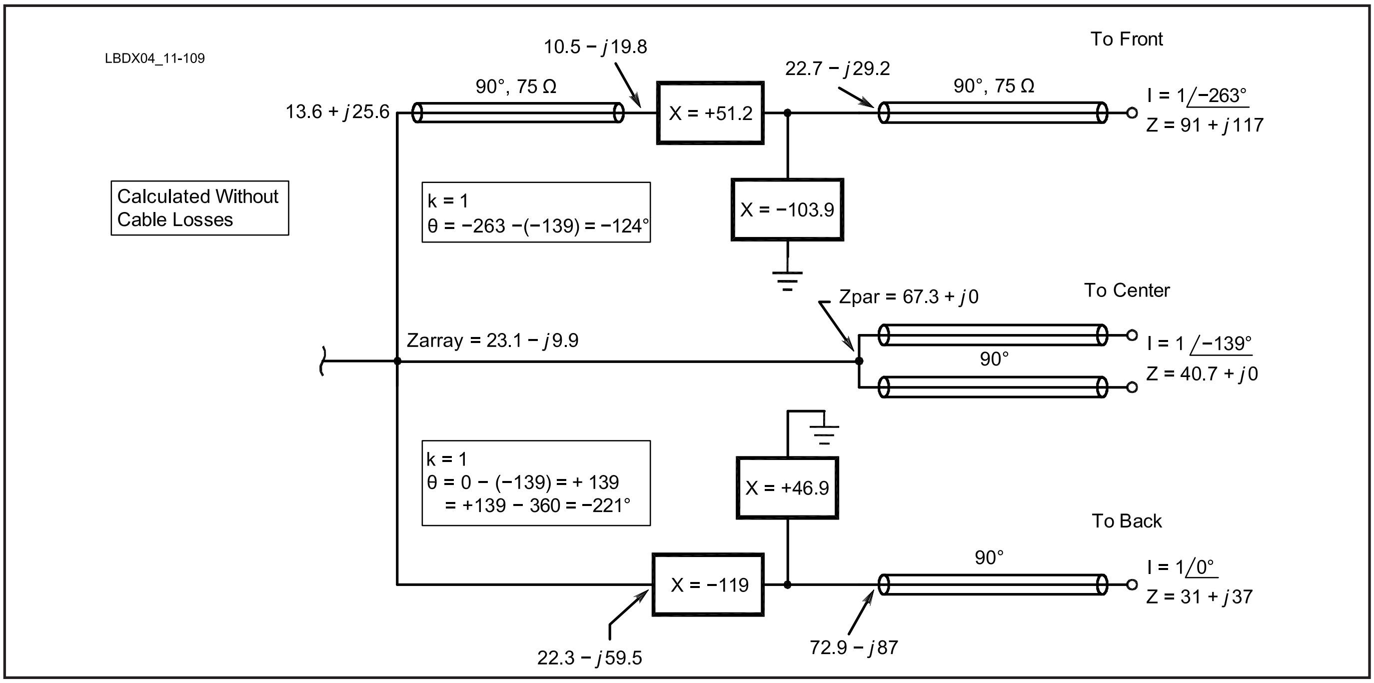



**Fig 11-109—Lahlum/Lewallen feed network for the Four-Square with bent half-wave dipoles shown in Fig 11-108A.**

central support tower are:

Z1 = 31 + *j* 37
Z2 = Z3 = 40.7 + *j* 0 Ω
Z4 = 91 + *j* 117

Fig 11-108B and 108C show the horizontal and vertical radiation patterns. For comparison, the vertical pattern of a single vertical is also included. **Fig 11-109** shows a feed method using the Lahlum/Lewallen feed system. The array shows a fair amount of high-angle radiation, due to the horizontal radiation component originated by the sloping wires. That's also why the RDF is pretty poor (8.06 dB vs 11.04 dB for the K8UR-antenna described in Section 7.3 below).

## 7.3. The K8UR Sloping-Dipole Four-Square Array

D. C. Mitchell, K8UR, described his 4-element sloping array in Ref 975. He uses half-wave sloping dipoles (sometimes called *slopers*) where the bottom half is sloped back toward the tower. This eliminates much of the high-angle radiation, as most of the horizontal component is now canceled due to the folding of the elements. The four feed points form a square measuring λ/4 on each side. See **Fig 11-110A**.

The original design was fed in quadrature. We now know that much better directivity can be obtained with phasing angles that are much larger than 90°. Modeling turned out a design that has excellent properties:
Feed currents:

I1 = 1 /0° A
I2 = I3 = 1 /–137° A
I4 = 1 /–263° A

Gain: 6.3 dBi (over average ground)
3-dB beamwidth: 88.4° (120° if quadrature fed!)
RDF = 11.07 dB
DMF = 22.7 dB

The feed-point impedances calculated with a grounded 58-meter support tower are:
Z1 = 24.5 + *j* 83 Ω
Z2 = Z3 = 36.3 Ω
Z4 = –17.8 + *j* 12 Ω

**Fig 11-111** shows the radiation patterns. For comparison, the vertical pattern of a single quarter-wave vertical is also included in B. Mitchell used a Collins-type, quadrature-feed system, although he replaced the 180° phasing line with a hybrid-type network, taking care of the required 180° phase shift. These hybrid couplers are now commercialized by Comtek Systems. Fig 11-111 shows a possible Lahlum/Lewallen feed system for this array.

Mike Greenway, K4PI, developed an innovative way for switching the dipoles of his K8UR-type array from the phone end of the band to the CW end of the band. See **Fig 11-112**.

If you have a tall tower, a Four-Square sloping array à la K8UR is a good way to go, with a clean pattern, excellent gain, good RDF and DMF because there is no high-angle radiation. This array lends itself to a feed design according to the cross-fire principle (see Section 3.4.4 and 4.7.4), where we end up with exactly 75 Ω for $Z_{par}$, using the two feed lines to the center elements in parallel (see **Fig 11-113**).

I inserted a 1:1 (180° phase-shift) transformer in the center branch. The required phase shift vs the front element (the element being fed directly) is –137 – (–263) = +126°. The coax has to take care of: –180 + 126 = –54°. The back element needs to be fed with a phase angle of +263° vs the front element, and that is the same as +263 – 360 = –97°.

John Brosnahan, WØUN, pointed out that because of the ends of the dipoles being close together (near the tower), mutual coupling is very high. This also shows in the high negative impedance of the back element, which means there is a lot of power coupled into this element by mutual coupling, power, that is fed back into the network. This results in a high-

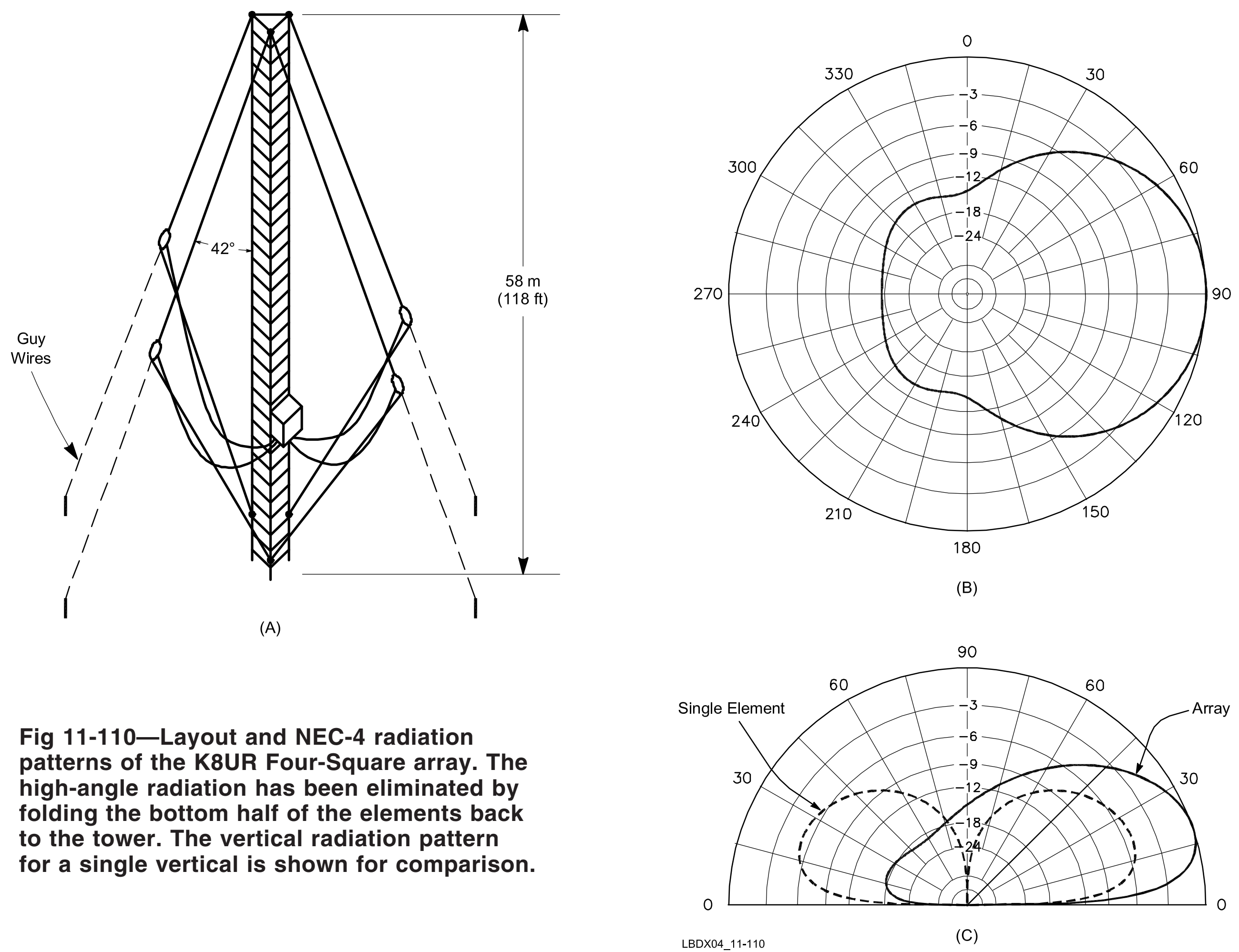



**Fig 11-110—Layout and NEC-4 radiation patterns of the K8UR Four-Square array. The high-angle radiation has been eliminated by folding the bottom half of the elements back to the tower. The vertical radiation pattern for a single vertical is shown for comparison.**

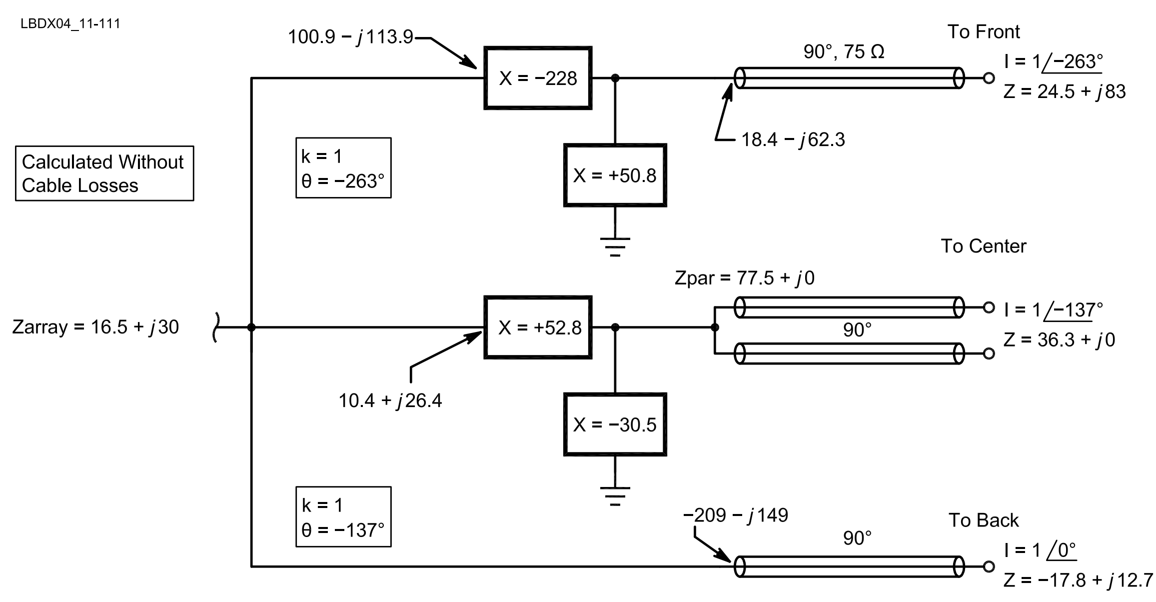



**Fig 11-111—An optimized Lahlum/Lewallen feed system for the K8UR array.**

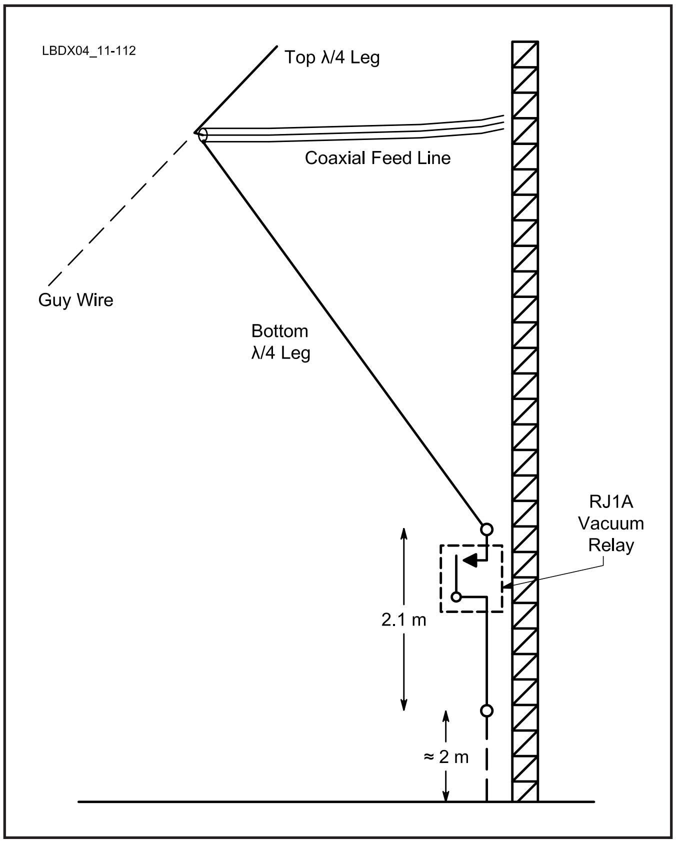



**Fig 11-112—Method for switching a K8UR-style dipole from 3.8 to 3.5 MHz. The bottom end of the dipole is lengthened with a piece of wire (which can be called capacitive loading) to lower its resonant frequency to 3.5 MHz. K4PI uses a 2.1-meter long wire, spaced about 0.5 meters from the tower. The relay must be able to withstand high voltage. Mike uses RJ1A vacuum relays.**

Q system and a narrow SWR and pattern bandwidth.

### 7.3.1. Conclusion, K8UR array

If you care about bandwidth, and are not just a CW operator, I would opt for two switchable L-networks in the Lahlum/Lewallen feed system for the K8UR element configuration. I'd adjust each for peak performance in the CW and SSB portions of the 80-meter band instead of using all sorts of tricks to try to improve operational bandwidth. But I find it interesting to analyze these cases, one by one, as they do improve our understanding in these matters.

## 7.4. The Four-Square Array with Sloping Quarter-Wave Verticals

If you cannot run long catenary cables to support wire vertical elements, you can stick some insulating booms made of fiberglass or aluminum broken up with insulators at the top of your tower. See **Fig 11-114**.

### 7.4.1. Array data

F = 3.75 MHz
Length sloping wires: 19.86 meters
Side of square: 20 meters
Feed currents:
I1 = 1 /0° A
I2 = I3 = 1 /–120° A
I4 = 0.9 /–235° A
Gain: 6.04 dBi
3-dB beamwidth: 87°
RDF: 10.55 dB
DMF: 22.4 dB
Feed-point impedances:
Z1 = 24.5 + *j* 60 Ω
Z2 = Z4 = 25.6 + *j* 5 Ω
Z3 = –7 + *j* 5 Ω

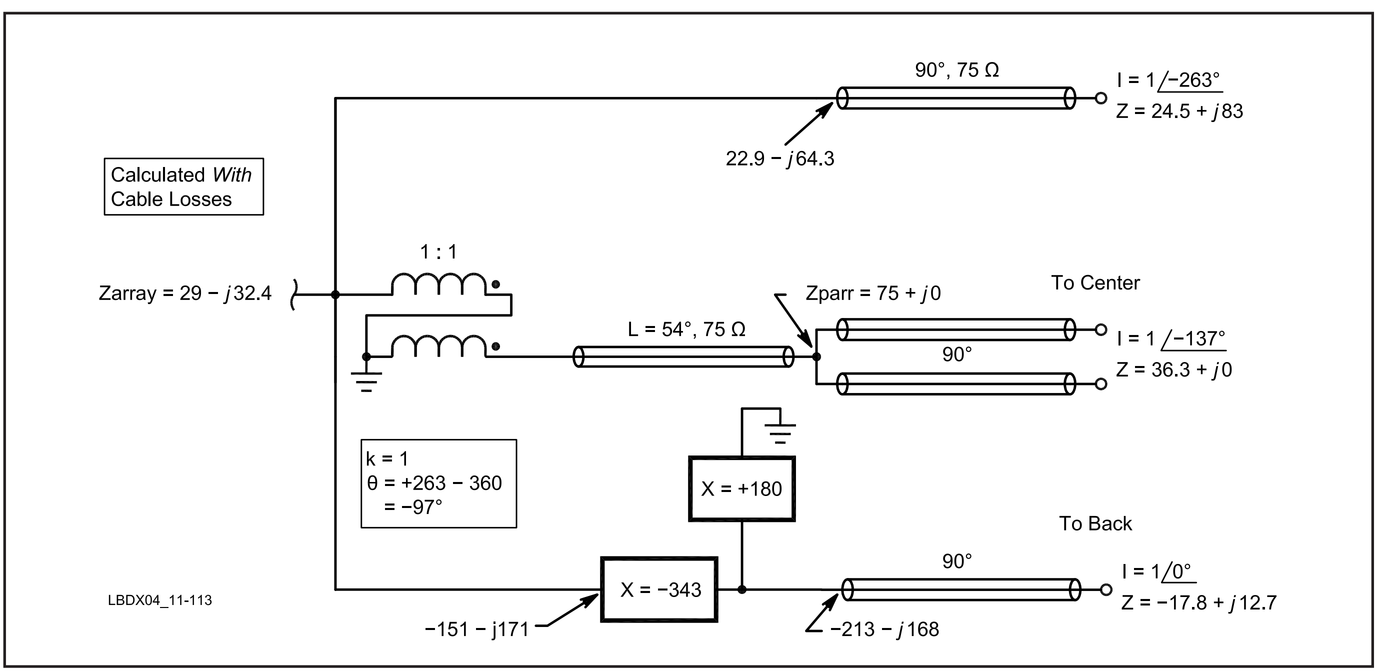



**Fig 11-113—Cross-fire principle applied to the K8UR array with optimized phasing.**

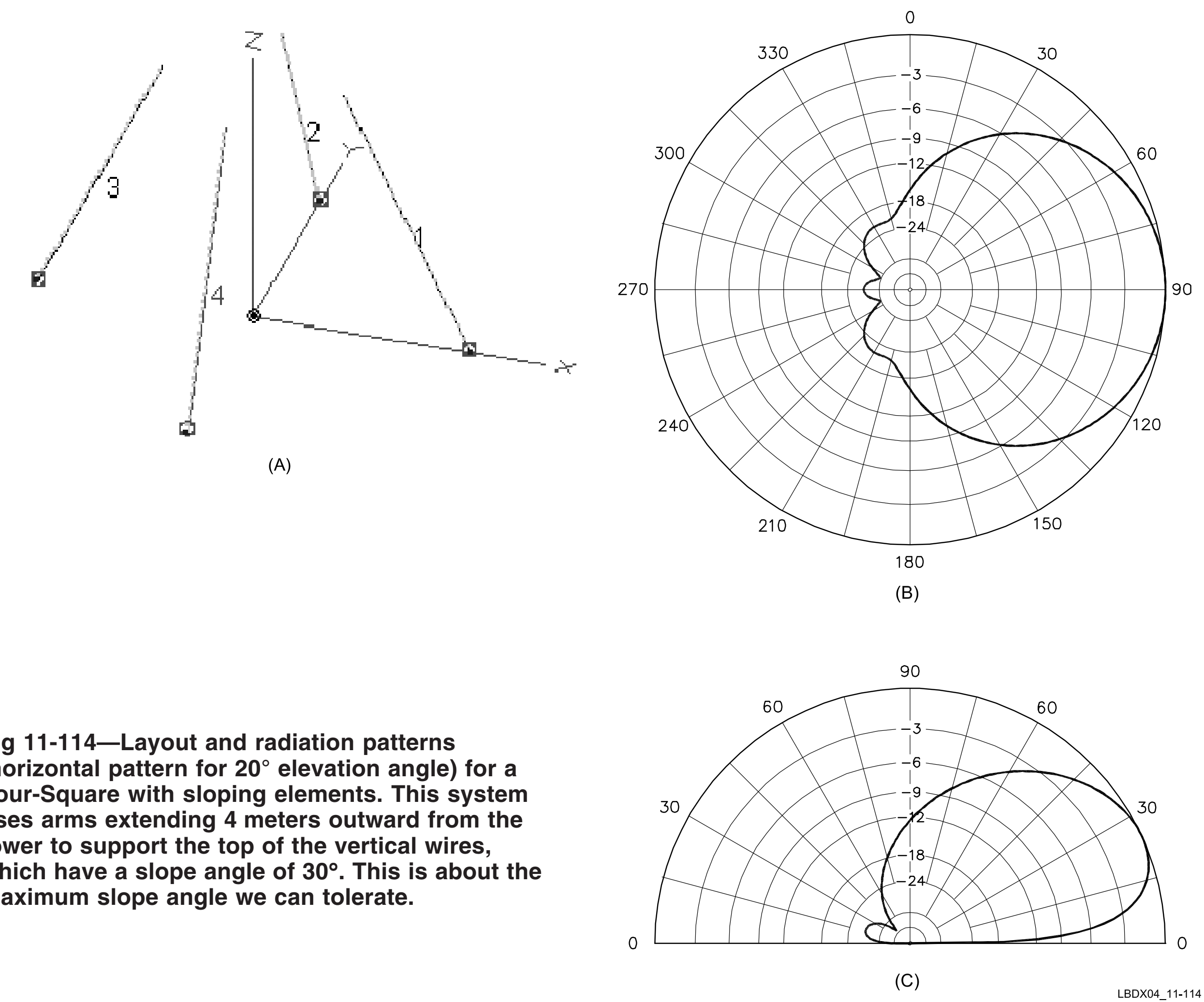



**Fig 11-114—Layout and radiation patterns (horizontal pattern for 20° elevation angle) for a Four-Square with sloping elements. This system uses arms extending 4 meters outward from the tower to support the top of the vertical wires, which have a slope angle of 30°. This is about the maximum slope angle we can tolerate.**

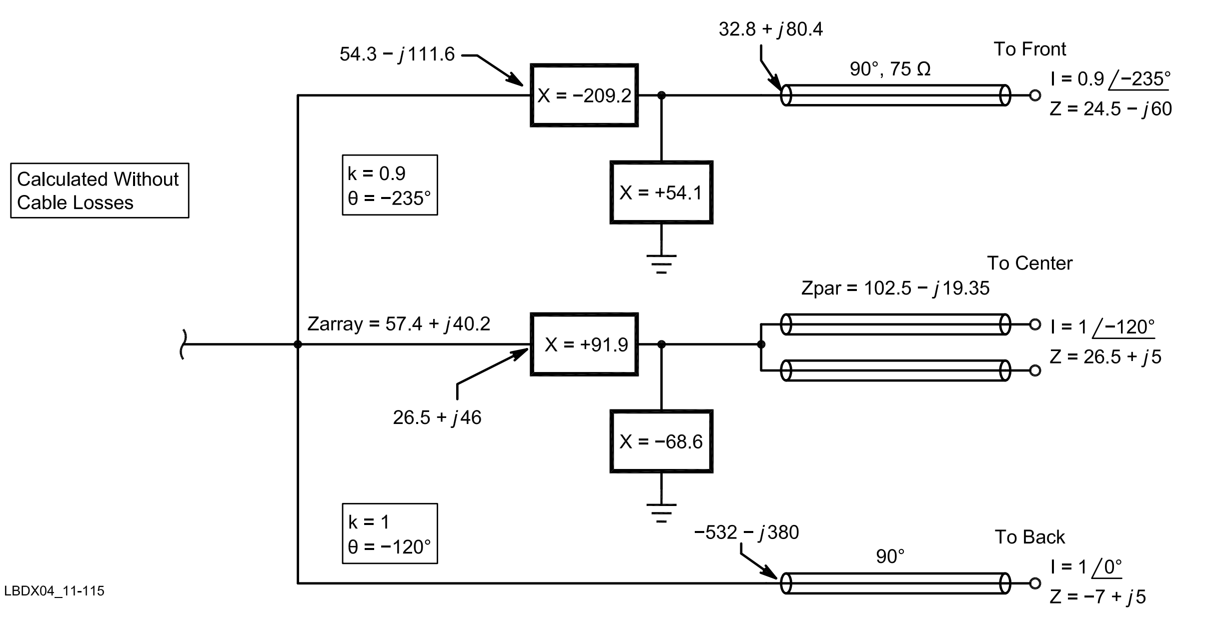



**Fig 1-115—One possible Lahlum/Lewallen feed method for a Four-Square with sloping elements shown in Fig 11-114.**

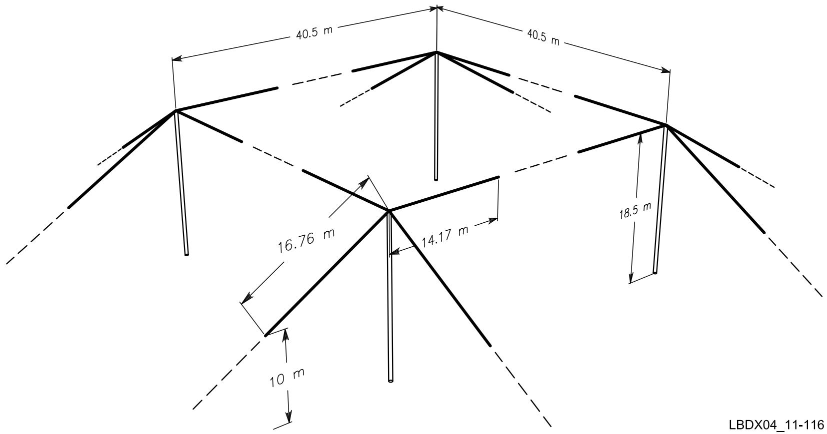



**Fig 11-116—This reduced-height Four-Square measures λ/4 along the side of the square and it uses λ/8-high elements. The loading wires are symmetrical and don't contribute high-angle radiation. Some fine tuning of the resonant frequency of the elements can be done by changing the height of the ends of the sloping loading wires (nominally 10 meters high).**

This array is very attractive if you have one central support and if you cannot run long catenary sloping wires to support the wire verticals. Make sure the central support does not upset the radiation pattern. If it is grounded, it may be too close to resonance and you may have to decouple it (see Chapter 7). **Fig 11-115** shows a Lahlum/Lewallen feed system for this array.

If you want to operate this array on both the CW and the phone end of the band, it is best first to model the antenna at 3.51 MHz. For that case the sloping vertical wires should be about 20.9 meters long (assuming 2 mm OD copper conductors). On 3.775 MHz the elements are too long and need to be shortened by a series capacitor of about 600 pF. A relay can switch the capacitor in and out of the circuit. At the same time the current-forcing feed line can be lengthened on 3.5 MHz to maintain its proper (λ/4) length. See also Fig 11-96.

It is clear that the same configuration can be used in 160 meters, where everything will be approximately twice the size.

## 7.5. An Attractive 160-Meter Four-Square With 18.5-Meter Tall Verticals

One does *not* need to use full-size λ/4 long vertical radiators to build a high-performance array. Using short verticals has a few consequences though:

1. Lower feed point impedances
2. Meaning that an excellent radial system is even more important
3. A low-loss top-loading configuration is essential

Top loading means wire loading. While inverted-L elements can be used, they will radiate a lot at high angles and completely destroy the RDF and DMF directivity figures of the array. Top-loading wires need to be arranged in such a way that far-field cancellation occurs for all radiation from the loading wires.

The design shown in **Fig 11-116** is a good example of such a design. Compared to a Four-Square with full-size (39-meter long) elements, the quadrature fed version of this array with 18.5-meter long elements (46% of full-size) gives up only 0.5 dB in gain, assuming an identical radial system is used (with 2-Ω equivalent loss resistance).

### 7.5.1. Quadrature fed

Dimension of square side: λ/4
Elements: 18.5 meters tall with top loading wires.
Feed currents:
I1 = 1 /–180° A (front element)
I2 = I4 = 1 /–90° A(center elements)
I3 = 1 /0° A (back element)
Gain: 6.10 dBi (Average Ground) and 7.73 dBi (Very Good Ground)
3-dB beamwidth: 100°
RDF = 10.20 dB
DMF = 19.1 dB
Feed-point impedances:
Z1 = 27.4 + $j$ 36.2 Ω
Z2 = Z4 = 19.5 – $j$ 1.2 Ω
Z3 = –1.8 – $j$ 1.4 Ω

### 7.5.2. Feeding the array

As this array is quadrature-fed you can feed it with a hybrid coupler (eg, Comtek). A Lahlum/Lewallen feed system is shown in **Fig 11-117**.

### 7.5.3. Optimized feed, Short Four-Square

We can get slightly more gain and better directivity with the WA3FET feed-current values. **Fig 11-118** shows the radiation patterns for this array.
Dimension of square side: λ/4
Feed currents:
I1 = 0.872 /–218° A (front element)
I2 = I4 = 0.9 /–111° A(center elements)
I3 = 1 /+ 0° A (back element)
Gain: 6.58 dBi (Average Ground) and 8.22 dBi (Very Good Ground)
3-dB beamwidth: 87°
RDF = 11.09 dB
DMF = 24.8 dB
Feed-point impedances:
Z1 = 16.5 + $j$ 29.9 Ω
Z2 = Z4 = 14.3 – $j$ 1.5 Ω
Z3 = 0.9 – $j$ 0.2 Ω

In this configuration you must use an L-network feed system (Lahlum/ Lewallen). **Fig 11-119** shows one of the possible solutions.

## 8. RADIAL SYSTEMS FOR ARRAYS

## 8.1. Buried Radials

Radials of the elements of an array cross each other. It is

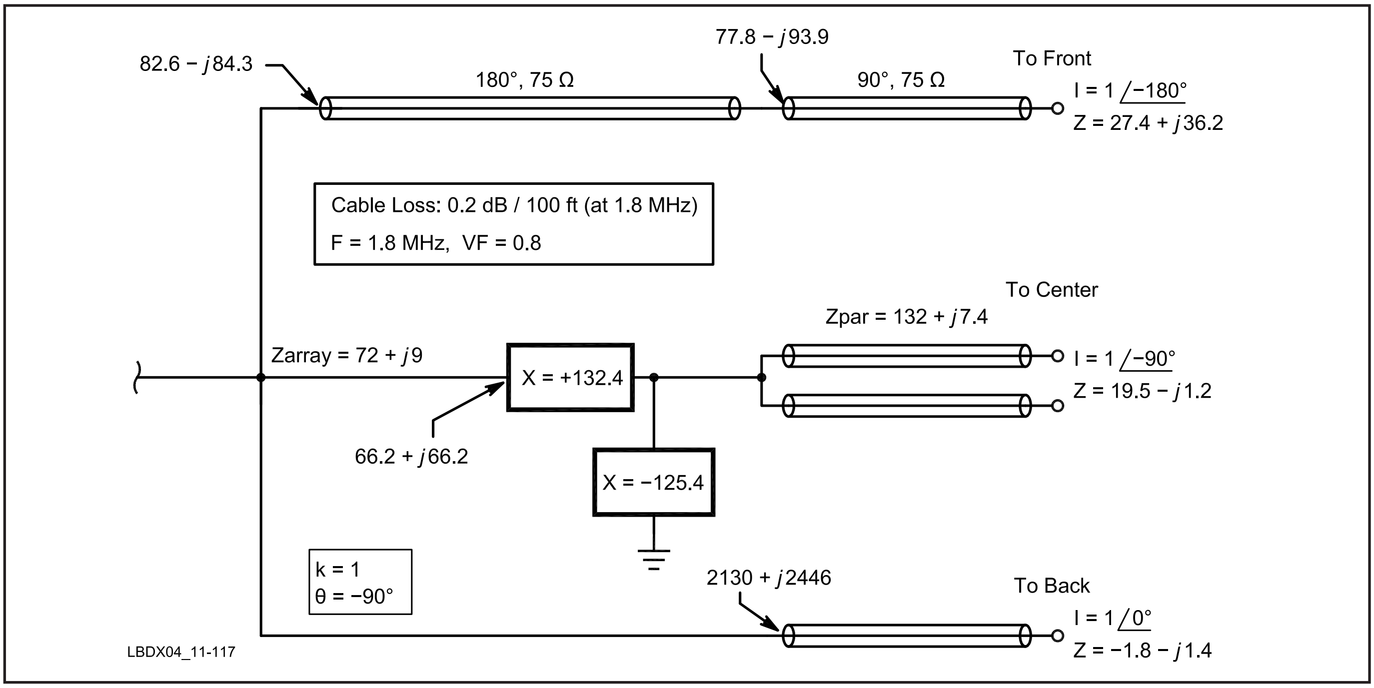



**Fig 11-117—Lahlum/Lewallen feed system for the quadrature-fed Four-Square with λ/8 top loaded elements.**

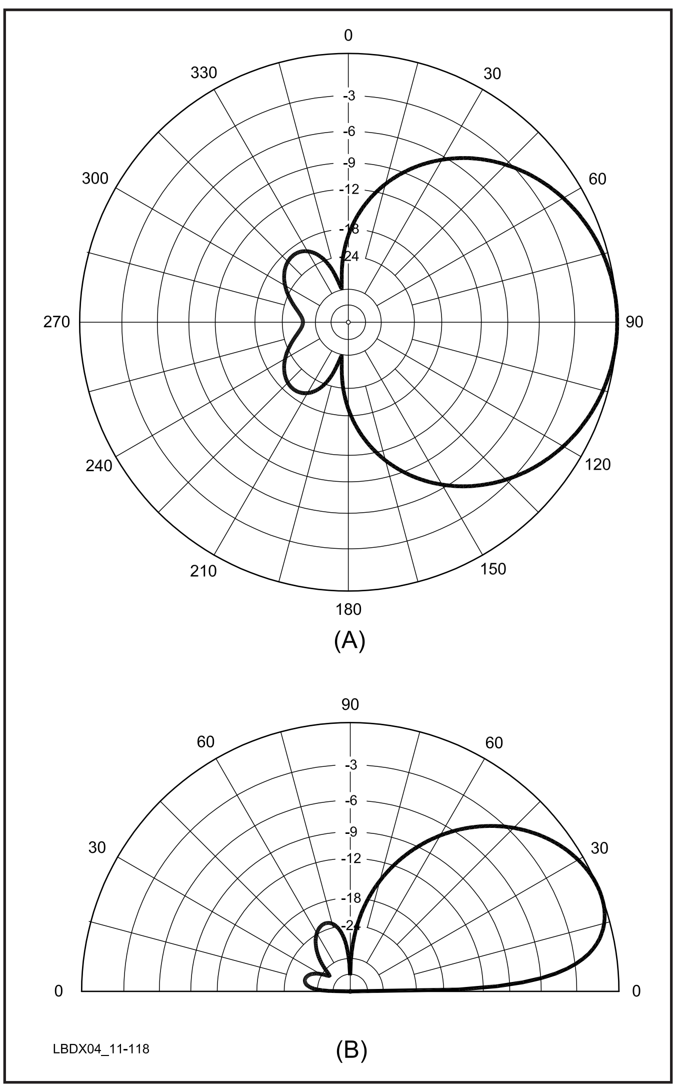



**Fig 11-118—Radiation patterns (horizontal for 20° elevation angle) for the optimized 160-meter Four-Square with λ/8-long top loaded elements.**

standard procedure to install a bus (AWG #6 or #8 copper strip) half way between the elements and to connect the radials to this "bus." The radials can be any wire size, if many are used. The size will be dictated more by mechanical strength than current-carrying capability. **Fig 11-120** shows various typical radial layouts for a 2-element cardioid array, a 3-element triangular array and for the classic Four-Square array.

## 8.2. Elevated Radials

When a large number of elevated radials are used on each of the elements of an array, these radials become non-resonant, and they can be connected to a bus system in exactly the same way as shown in **Fig 11-120**.

When only a few radials are used (typically 1 to 4 radials), the situation is very different. In this case the radials can couple heavily with adjacent (especially parallel) radials from other elements and can upset the directivity of the array, and create uncontrolled and unwanted high-angle radiation from the elevated radials. **Fig 11-121** shows a few possible layouts that try to minimize the coupling.

## 9. CONCLUSION

Now that we have powerful modeling programs available (eg, *EZNEC*), I would like to encourage everyone to try to develop an array that fits his or her own requirements. The dBi gain is *not* the gain you should expect if you change from a single vertical to an array. Over average ground (dielectric constant = 13, conductivity = 5 mS) a single element has a gain in the order of 1 dBi, so if your modeling program tell you the array has 6 dBi gain, expect about 5 dB gain over a single vertical.

During your modeling sessions, make sure you use the

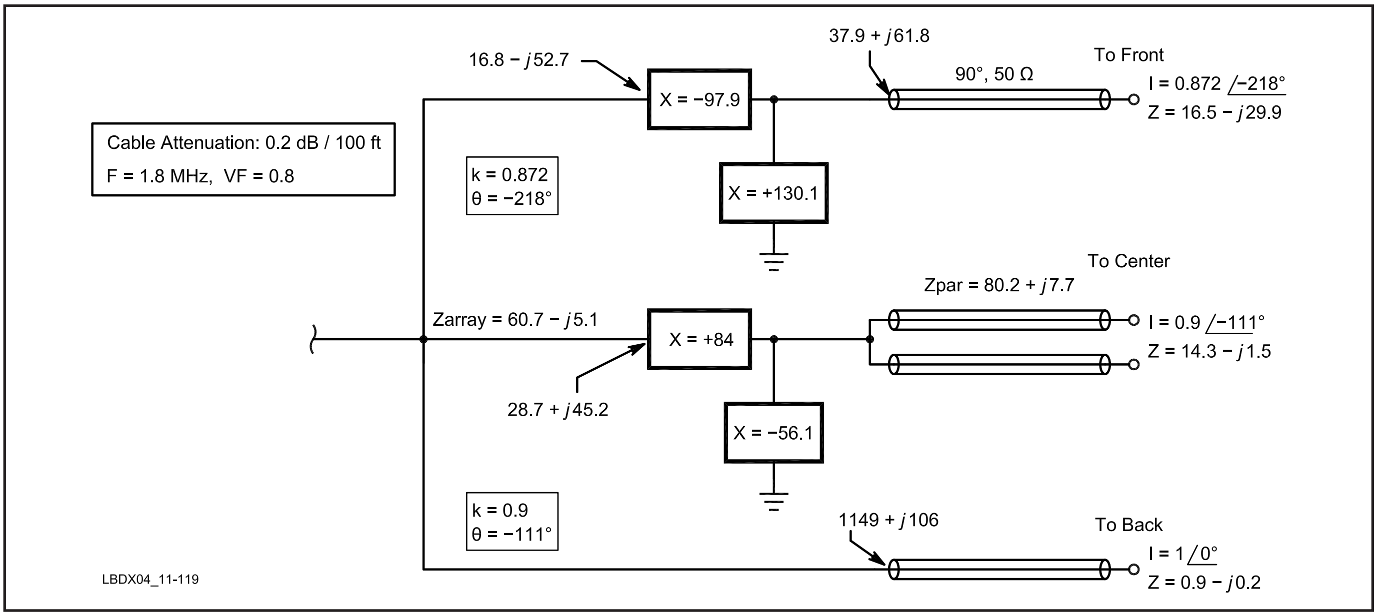



**Fig 11-119—Lahlum/Lewallen feed system for the optimized short-element Four-Square for 160 meters. This time 50-Ω feed lines turned out to be the better choice.**

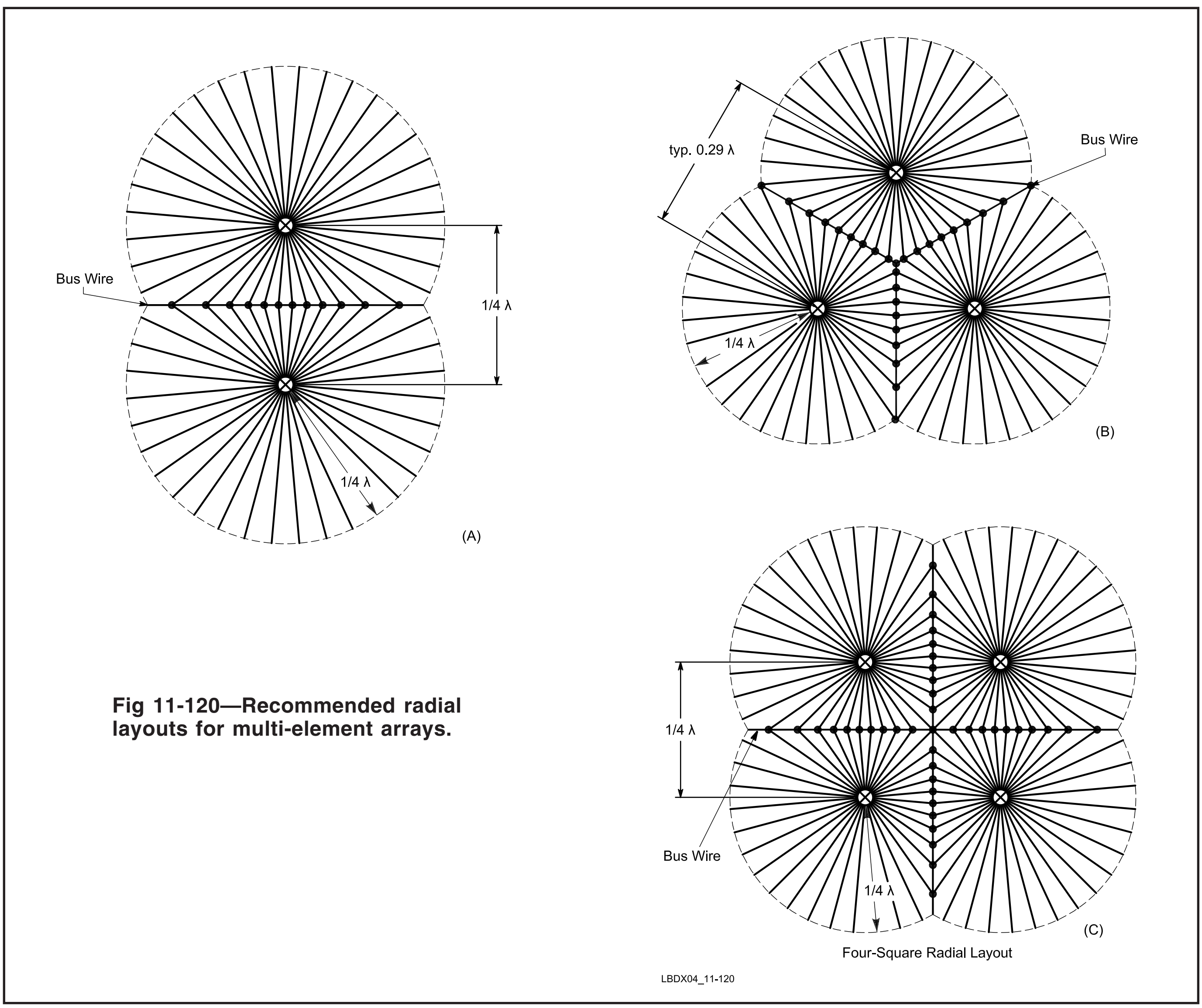



**Fig 11-120—Recommended radial layouts for multi-element arrays.**

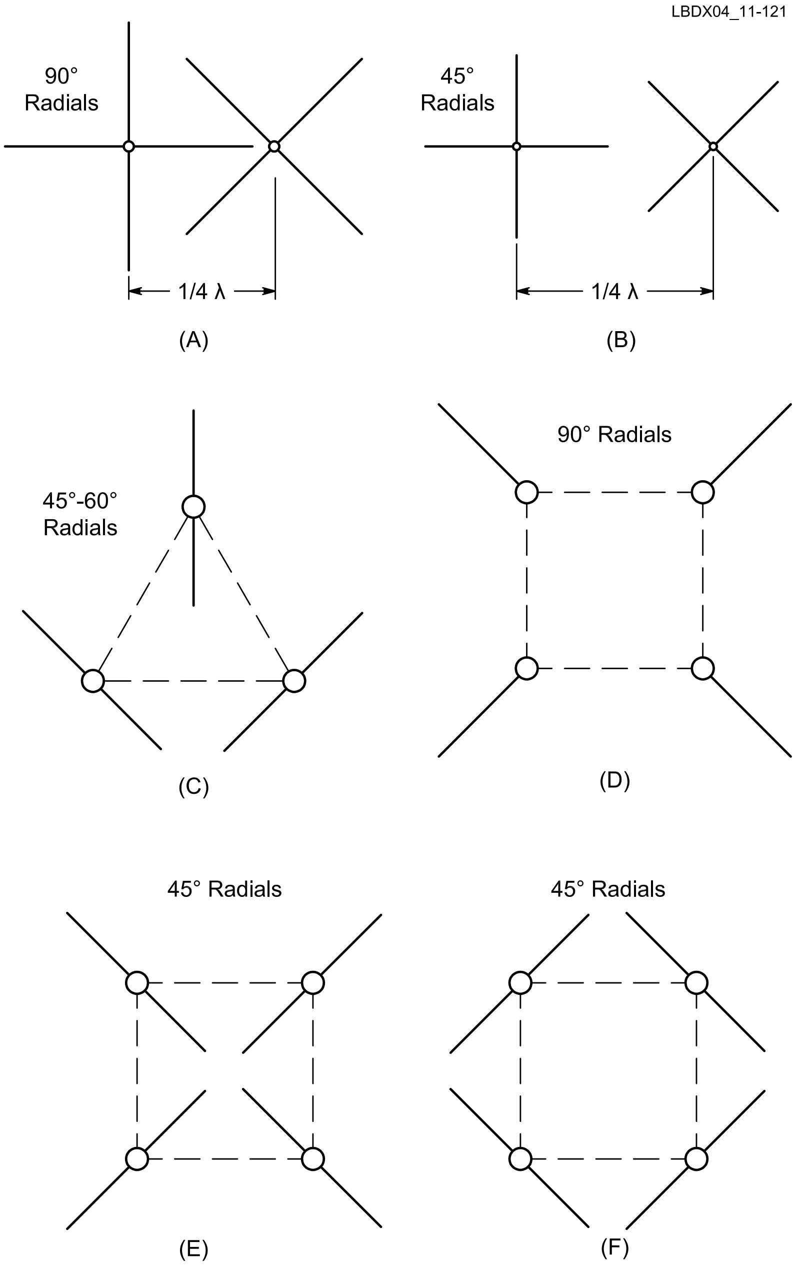



**Fig 11-121—Recommended elevated radial layout schemes.**

same ground quality specifications in all circumstances. If you want to model the antenna for designing your own feed system, don't forget to include the equivalent loss resistance for the-less-than perfect ground radial system.

Let me warn the readers once again: Antenna modeling is one thing and practical antenna design can be a totally different thing. You can model complex arrays with staggering characteristics, but which may be difficult to build and which might not a project for a newcomer with little technical expertise to attempt to build.

Don't forget you have to feed the array, by practical means, with real-world feed lines. After having done your initial calculations without taking cable losses into account, give it a final go with losses. Especially if extreme impedances (high or low) are involved and relatively long coax runs (eg, 270° feed lines) you may be surprised how much the difference between the ideal world (no losses) and the real world can be!

If you have to compromise a little in order to be able to use a much easier-to-make feed-system (eg, quadrature fed versus the requirement for exotic phase angles or current magnitudes), I would advise the compromise, unless you have the required measuring setup.

A simple test system developed by W1MK makes it possible to tune each L-network feed system for top-notch performance. There no longer is a need for vector volt-meters. You can now get top performance just using very affordable test instruments that you can build yourself. Go for it!

If designing and building your own feed system does not attract you, know that you can nowadays buy systems using the Lahlum/Lewallen approach commercially (from Array Solutions).